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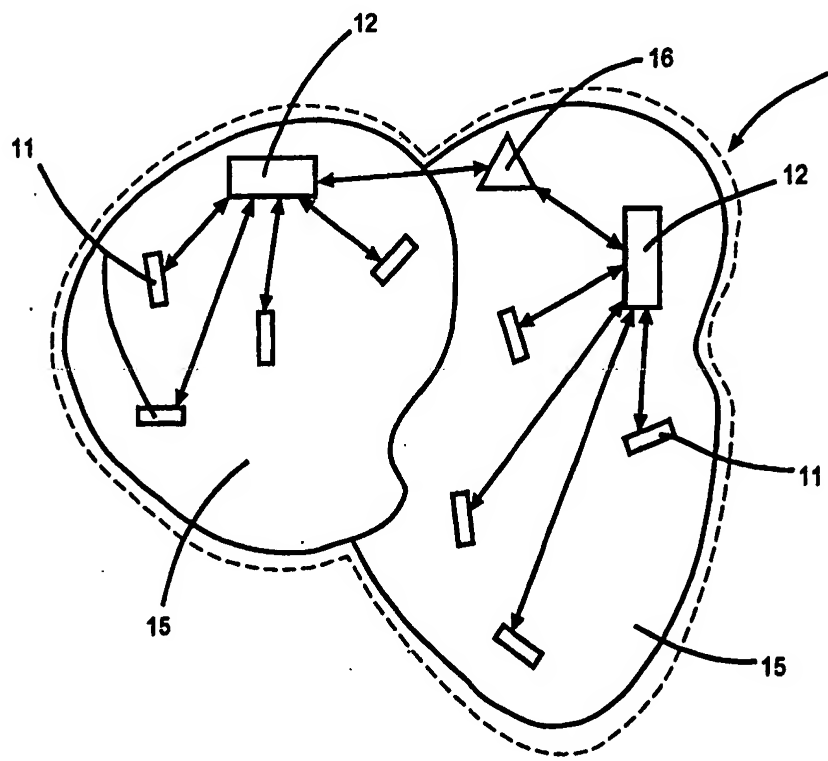
(54) Title: CIRCUITRY AND METHOD FOR DEMODULATING CODE DIVISION MULTIPLE ACCESS (CDMA) SIGNALS

(57) Abstract

A method and apparatus is provided for demodulating direct sequences, spread spectrum CDMA signals in the presence of unknown spread-spectrum multiuser interference and additive white Gaussian noise, using only the available received signal and the predetermined, prestored spreading code assigned to user of interest. The method includes the steps of sampling the incoming channel signal at the chip rate to produce an incoming sampled signal, supplying the incoming sampled signal to filter means having tap coefficients corresponding to an auxiliary vector, the filter means having an output representing an estimate of the transmitted symbol, providing the output to a sign determining means and determining the sign of said output to thereby generate the detected symbol.



CIRCUITRY AND METHOD FOR DEMODULATING CODE DIVISION MULTIPLE ACCESS (CDMA) SIGNALS

BACKGROUND OF THE INVENTION

1. Field of the Invention

The present invention relates to communication systems which employ spread-spectrum signals and, more particularly, to receivers for detecting spread-spectrum signals in a Direct Sequence (DS) CDMA communication system.

2. Background of the Invention

The wireless telecommunications industry provides the capability of telephony to a moving public through cellular phones. CDMA is a military-based digital communication system and a digital standard for cellular phone systems. A special case of the CDMA system is a spread-spectrum system, i.e. spread spectrum CDMA or SS/CDMA. In a spread-spectrum system, a modulation technique is utilized in which a transmitted signal is spread over a wide frequency band within the communication channel.

One type of spread-spectrum communication technique, direct sequence modulation, relates particularly to the present invention. In direct sequence (DS)modulation, a carrier signal is modulated by a digital code sequence whose bit rate is much higher than the information signal bandwidth. In direct sequence systems communication between two communication units is accomplished by spreading each transmitted signal over the frequency band of the communication channel with a unique user spreading code, hereinafter referred to as a signature or code. As a result, transmitted signals are in the same frequency band of the communication channel and are separated only by the unique use spreading codes or signatures.

Information (i.e. the message signal consisting of voice and/or data) can be embedded in the direct sequence spread-spectrum signal by several methods. One method is to add the information to the spreading code before it is used for spreading modulation. Alternatively, the information or message signal may be used to modulate a carrier after spreading it.

Particular transmitted signals can be retrieved from the communication channel by despreading a signal representative of the sum of signals in the communication channel with a user spreading code related to the particular transmitted signal which is to be retrieved from the communication channel.

Preferably, the user spreading codes are orthogonal to one another such that when the received signal is correlated with a particular user spreading code, only the desired user signal related to the particular spreading code is preserved, while the other signals for all of the other users are eliminated. In many cases the spreading codes are not orthogonal. This may occur by design or may exist due to the transmission characteristics of the channel.

Several different spreading codes are known which can be used to separate data signals from one another in a DS CDMA communication system. These spreading codes include but are not limited to pseudo noise (PN) codes and Walsh codes.

In a mobile cellular phone system, there are problems inherent in designing DS/SS CDMA receivers for either the mobile user or the base station with which it communicates. The base station must simultaneously detect and recover all signals of the known mobile users in its domain. Given that the optimum multiuser detector exhibits unrealizable complexity, i.e. exponential in the number of users and system processing gain, suboptimal linear-complexity solutions were sought. Some proposals

include the decorrelating receiver, multistage architectures and decision feedback detecotrs.

The mobile users, however, have a more difficult problem. Knowing only its own signature, a mobile user's receiver must detect its own information bits in the presence of unknown spread spectrum, or multiple access (MAI) interference, and additive white Gaussian (AWG) channel noise.

Increasing the challenge of designing the mobile receiver is the fact that it must be of a desirable size and weight, i.e. lighter and smaller than would be acceptable for a base station receiver.

Demodulating a DS/SS CDMA signal in the presence of multiple access interference has been previously addressed in the prior art. One solution, the "matched filter (MF) solution exhibits performance degradation in the presence of one or more high power interferers, known as the "near-far" problem. This characteristic of MF receivers requires that they be used with some form of stringent, and costly, power control.

U.S. Patent No. 5,345,472 to Lee discloses a DS/SS CDMA receiver for adaptively decoding DS/SS communication signals. In this system the CDMA transmitter transmits a training bit sequence and the receivers adaptively determine, based on the training sequence, the despreading codes. This is accomplished by converging or minimizing the error between the received training bit sequence and the reference bit sequences.

While this system allows all users to communicate with each other over a channel without requiring knowledge of system parameters, it disadvantageously requires transmission of a separate training sequence which places a concomitant processing burden on the receiver.

U.S. Patent No. 5,343,496 to Honig et al. discloses circuitry and concomitant methodology for demodulating DS/SS CDMA channel signals using multiple samples per transmitted symbol and a minimum mean-square error

criterion to suppress interference. Honig et al, propose to minimize the mean-square error (MSE) between the output of the filter and the desired information bit. However , this technique has the disadvantage of requiring a separate training sequence.

SUMMARY OF THE INVENTION

These deficiencies as well as other shortcomings and limitations of the prior art are obviated, in accordance with the present invention, by circuitry and methods for demodulating CDMA signals by parametizing a linear detector with respect to a single scalar parameter.

Broadly, in accordance with a first aspect of the invention, a CDMA receiver incorporates a circuit for demodulating a received signal wherein the received signal includes a transmitted symbol, unwanted interference and noise. The circuit includes means for obtaining one or more discrete samples of the received signal and means, responsive to the means for obtaining one or more discrete samples, for filtering out the unwanted interference and noise. The filtering out means includes multiplying means having at least two sets of inputs, at least one set being determined by the relationship between the parameters S_0 ,k, and G. In one embodiment of the invention the relationship is $S_0 - kG$. In a second embodiment the relationship is $G + kS_0$.

It is a further object of the present invention to provide a method for demodulating a received signal wherein the received signal includes a transmitted symbol and unwanted interference and noise, the method including the steps of: a) obtaining one or more discrete samples of the received signal; b) filtering out the unwanted interference and noise from the discrete samples wherein the filtering out step includes the step of multiplying at least two sets of inputs, at least one set being determined by the relationship between the parameters S_0 ,k, and G. The method may further include the steps of

providing the output of step b to sign determining means, determining the sign of the filtered output and generating a representation of the transmitted symbol based on the sign determined in the previous step.

BRIEF DESCRIPTION OF THE DRAWINGS

For a further understanding of the invention, as well as its characterizing features, reference should now be made to the accompanying drawings wherein:

FIG. 1 is a pictorial diagram of a communication system of interest in accordance with the present invention.

FIG. 2 is a pictorial diagram showing a base station, a representative mobile user, and a communication channel therebetween.

FIG. 3 depicts the relationship between a data symbol stream representing data to be transmitted and a signature encoded chip bit stream propagated in correspondence to the data symbol stream.

FIG. 4 is a block diagram of a CDMA receiver in accordance with the present invention.

FIG. 5 is a detailed block diagram of a demodulator according to the present invention.

FIG. 6 shows a comparison of the bit error rates per number of data samples of the detector of the present invention with a detector of the Minimum Output Energy (MOE) type.

FIG. 7 shows a comparison of the detector of the present invention with detectors of the matched filter, decorrelator and MOE types, as well as with an ideal MOE detector on the basis of bit error rate as a function of the SNR of the user of interest in the presence of strong interferers.

FIG. 8 shows a comparison of the detector of the present invention with detectors of the matched filter, decorrelator, MOE types, as well as with the ideal minimum output energy detector on the basis of bit error rate as

a function of the SNR of the user of interest in the presence of weak interferers.

FIG. 9 shows a Weiner reconstruction filter embodiment of the present invention.

DETAILED DESCRIPTION OF THE INVENTION

The detector of the present invention provides for the detection of binary antipodal signals, such as those employed in DS/SS CDMA communications systems, in the presence of unknown spread-spectrum multiuser interference and additive white Gaussian noise (AWGN). The detector of the present invention reconstructs the transmitted symbol from the received composite signal using only the predetermined spreading code(referred to herein as the signature, or code) assigned to the user of interest, i.e. the user to whom the information is being transmitted.

FIG. 1 depicts a typical cellular telephony system 10. In system 10 a geographical area, e.g., a metropolitan area, is divided into several smaller, contiguous radio coverage areas, or cells 15. Cells 15 are served by a series of fixed radio stations, or base stations 12. Base stations 12 are connected to and controlled by a mobile services switching center (MSC) 16. MSC 16, in turn, is connected to the landline(wireline) public switched telephone network (PSTN) (not shown).

The telephone users, or mobile subscribers 11 in the cellular radio system are provided with portable (hand-held), transportable (hand-carried) or mobile (car-mounted) telephone units (mobile stations) which communicate voice and/or data with each other through a nearby base station 12. MSC 16 switches calls between and among wireline and mobile subscribers 11, controls signaling to the mobile subscribers 11, compiles billing statistics, and provides for the operation, maintenance and testing of the system 10.

FIG. 2 is a pictorial representation of a representative base station 12 and a representative mobile subscriber 11 of the system shown in FIG. 1. Communication system 10 can be considered to comprise generally a plurality of sources, or transmitters 120, of which two representative examples are shown in FIG. 2, and a plurality of receivers 122 of which two representatives are also shown in FIG. 2.

In the wireless telecommunication system contemplated by the present invention, a receiver may be associated with either a base station or a mobile unit. Likewise a transmitter may be associated with either a base station or a mobile unit.

The sources, or transmitters 120 and receivers 122 are interconnected by a communications channel 130. Channel 130 propagates DS/SS CDMA signals produced by the sources 120. Each source 120 and each recievier 122 is associated with a pre selected CDMA signature sequence that corresponds to a particular subscriber and this siganture is used for both transmission and reception.

Each mobile receiver 122 and transmitter 120 knows of the unique spreading chip sequence, or signature, which will be used to transmit information between them prior to the start of a communication call. Thereafter, the receiver 122 decodes the DS-SS communication signal based on the known spreading chip sequence.

Each receiver, 122, demodulates the incoming DS SS CDMA channel signal by sampling the incoming channel signal at a rate corresponding to the processing gain of the CDMA channel signal to produce a sampled incoming signal and processes the discrete samples to finally detect the information bit of the user of interest. Processing gain is defined to be the ratio of the bandwidth of the spreading chip sequence, or chip rate, to the bandwidth, or bit rate, of the user signal.

Electromagnetic radiation propagating so as to carry information from a base station 12 to a mobile subscriber

11, as depicted by the arrow at 18 in FIG. 2, is hereinafter referred to as a downlink. Electromagnetic radiation propagating so as to carry information from a mobile subscriber 11 to a base station 12, as depicted by the arrow at 20 in FIG. 2, is hereinafter referred to as an uplink.

Referring now to FIG.3 there is shown at 210 a symbol stream representing data and/or information to be transmitted from a source to a receiver. FIG. 3 depicts three contiguous data symbols 203,204 and 205, in this case +1,-1,+1, respectively. Data symbols 203,204 and 205 are also referred to herein as information bits, or symbols.

Shown at 220 in FIG.3 is a rate increased transmitted pulse stream corresponding to the symbol stream at 210. As shown, a rate increased signature stream of (+1,-1,+1,+1,-1,-1) level pulses is propagated for each +1 (for example at 203 and 205) in the lower rate data symbol stream shown at 210. The inverse of this signature stream is propagated for each -1 (e.g., at 204) in the symbol stream.

The rate increased data stream shown at 220 of FIG.3 is the signature, or spreading chip sequence of the associated source. A frame is defined to correspond to a symbol having a duration T, shown at 201, and the time interval of each +1 or -1 level in the rate increased stream is designated the chip duration T_c (at 202). Processing gain is defined to be the ratio T/T_c. Each frame is composed of a fixed number of "chips" (also referred to as "spreading chips". Therefore, six +1 and -1 chips are propagated during each frame, or time T. Thus the signature associated with the representative example at 220 is the ordered set (+1,-1,+1,+1,-1,-1).

FIG. 4 shows, in block diagram form, a receiver 400 according to the present invention as it functions in the DS/SS CDMA system described generally above. Receiver 400 comprises signal receiving and pre conditioning means

including antenna 101, carrier demodulating means 102 and sampling means 103. Receiver 400 further comprises detector means 500 including filter means 105, sign determining means 107, multiplier input generating means 407 and means for storing a predetermined signature 410.

For purposes of this specification, and the following discussion, boldface variables shall be used to denote vectors in R^L, or to denote matrices. $E\{\cdot\}$ defines the statistical expectation operation, x^T is the transpose of x, and I or 1 denotes an identity matrix. The symbols $<\cdot,\cdot>$ and $\|\cdot\|$ denote inner product and the norm, respectively, both in R^L space unless otherwise specified.

In the system under consideration, K users, such as those depicted at 11 in FIG.1, transmit over an AWGN (Additive White Gaussian Noise) channel such as that represented in FIG. 2 at 130, to produce a received signal 109 at antenna means 101 as best shown in FIG.4.

Signal 109 contains the transmitted symbols which represent user data (a representative example of which is illustrated in FIG. 3). Signal 109 is applied to conventional preselecting and filtering means 102 which provides the initial receiver selectivity. Such means are well known in the art and need not be described here.

Preselecting and filtering means 102 may also include carrier demodulation means and may also perform low pass filtering. Such circuits are well known in the art. For example, preselecting and filtering means 102 may comprise a well known base band demodulator that demodulates the received signal 104 in accordance with the carrier modulation scheme used in the transmitter to provide a baseband signal at the output of carrier demodulating means 102.

The output of preselecting and filtering means 102 is continuous-time received signal r(t), represented at 104. Signal r(t) can be described mathematically as follows:

$$r(t) = \sum_i \sum_{k=0}^{K-1} \sqrt{E_k} b_k(i) s_k(t - iT) + n(t)$$

In this expression of r(t), with respect to the k-th user, E_k is the received energy, $b_k(i) \in \{-1,1\}$ is the i-th information bit, and $s_k(t)$ is the signature, also referred to herein as the spreading code. T is the transmitted symbol period as shown in FIG. 3 at 201. The filtered channel AWGN is n(t).

The signature of every user is composed of L spreading chips, as illustrated in FIG 3, and can be represented by

$$s_k(t) = \sum_{j=1}^{L} c_k(j) P_{T_c}[t - (j - 1) T_c]$$

where L also represents the system processing gain, which can be expressed as the ratio of T/T_c.

The assigned spreading pulses, or chips, for the k=th user are represented by: $c_k(j) \in \{-1,1\}$, j=1...,L. $P_{T_c}(t)$ is the spreading pulse, or chip, having duration $T_c=T/L$, as depicted in FIG. 3 at 202.

The signatures are assumed to be normalized such that they may have non zero cross correlation and such that

$$\int_0^T s_k^2(t)\,dt = 1, \ \forall k = 0,\ldots,K-1$$

Signal r(t) at 104 is applied to sampling means 103. Sampling means 103 can be any sampling means which performs chip ($P_{tc}(t)$)matched integration of the DS/SS CDMA signal r(t) at 104 and sampling at the chip rate. The result is dumped, or output at the end of each chip

interval of duration T_c producing an output such as represented at 110.

The design of such sampling means is well known in the art. For example, sampling means 103 may comprise a well known integrate and dump filter block wherein the received signal 104 is integrated and sampled at the chip rate and the results dumped at output 110 at the end of each chip interval of duration T_c.

As previously stated, sampling means 103 performs conventional chip matched filtering and sampling at the chip rate $1/T_c$, resulting in L collected samples at the output of sampling means 103 for every received frame, or chip interval L. These samples may be represented mathematically as elements of an L-dimensional vector as follows:

$$r[n] \triangleq \int_{(n-1)Tc}^{nTc} r(t)\, P_{T_c}[t-(n-1)T_c]\,dt,\ n = 1,\ldots,L$$

The R^L discrete time version of r(t) can be written in the form of an L-dimensional vector as:

$$\mathbf{r} = \sum_{k=0}^{K-1} \sqrt{E_k}\, b_k \mathbf{S}_k + \mathbf{n}$$

where S_k is the sampled signature of user k, i.e. a vector of dimension L, and where the random vector **n** is assumed to be white Gaussian noise with autocorrelation matrix

$$E\{\mathbf{n}^T\mathbf{n}\} = \sigma^2 \mathbf{1}_{LxL}$$

The Multiple Access Interference (MAI) term of the R^L discrete time version of r(t) can be written as:

$$I = \sum_{K=1}^{K-1} \sqrt{E_k} b_k S_k ,$$

Thus, if the user of interest is 0, and the predetermined spreading code transmitted for that user is S_0 then the available vector sample 110 present at the output of sampling means 103 can be described as follows:

$$r = \sqrt{E_0} b_0 S_0 + I + n$$

The sample at 110 is provided to both filter means 105 and multiplier input generating means 407 of detector means 500. Means for storing a predetermined signature 410 contains the sequence of chips representing the signature of the user of interest. Means for storing a predetermined signature 410 may be any electronic storage means utilized in storing digital signals for processing by a computer, microprocessor, or digital signal processor. These storage means include, but are not limited to, Random Access Memory (RAM) and Read Only Memory (ROM). The stored signature of the user of interest is provided to multiplier input generating means 407.

multiplier input generating means 407 utilizes the information from available vector sample 110 and the prestored signature to formulate multipliers, or tap weights to be provided at its output 415 to filter means 105.

Output 415 of multiplier input generating means 407 can be represented mathematically as a tap weight vector W where:

$$W^T = [w_1, w_1, \ldots, w_L]$$

Output 415 provides the tap coefficients, or multipliers, for filter means 105, which can be, in a preferred embodiment, a transversal, or finite impulse response (FIR) filter of a type well known in the art. Other appropriate filters, such as adaptive Wiener filters, preferably Weiner reconstruction filters, may be employed in the present invention and these remain within the scope of the invention.

Filter means 105 shifts, multiplies and sums the input vector samples at 110. The multipliers have two inputs. The mulitpliers have as one input the sequence of recieved signal samples. The other input is the sequence of filter coefficients described by the vector **W**. The output of filter means 105, for every frame of input data, is an estimation of the symbol of the user of interest, transmitted during that frame.

The output 106 of filter means 105 is provided to sign determining means 107. Sign determining means 107 determines the sign, positive or negative, of its input and provides an output representing information bits b_0 for the user of interest, i.e. user data.

Turning now to FIG. 5 there is shown a more detailed block diagram of a preferred embodiment of a detector 500 of a CDMA receiver constructed according to the teachings of the present invention. Detector 500 functions to produce at its output 106, a bit decision corresponding to the received symbol for each received frame.

As previously described, present at input 104 of sampling means 103 is the continuous time signal r(t). Continuous time signal r(t) is converted to a sampled data signal by sampling means 103. Sampling means 103, performs chip ($P_{tc}(t)$) matched integration of r(t) and sampling at the chip rate $1/T_C$ over time T to produce samples $r_1, r_2, \ldots r_{L-1}, r_L$ as depicted at 110. The samples are provided to one or more delay elements 501 having delay

Tc. Preferably, the number of delay elements is equal to L, i.e. the number of collected samples, or the processing gain.

The samples appearing at 502 are multiplied by the multipliers, or coefficients $(w_1 \ldots w_L)$ appearing at the inputs to multipliers 505. The resultant products 503 are summed in summer 510, with the resulting sum appearing at output 106 of summer 510. The resultant sum is provided to sign determining means 107.

Multiplier input generating means 407 utilizes the sampled signal r present at its input 415, and the prestored signature, or known spreading code, as represented at 410 assigned to the user of interest to accurately produce user data.

A comparison of the bit error rates per number of data samples of the detector of the present invention with a detector of the Minimum Output Energy (MOE) type is illustrated in FIG. 6.

FIG. 7 shows a comparison of the detector of the present invention with detectors of the matched filter, decorrelator and MOE types, as well as with an ideal MOE detector, on the basis of bit error rate as a function of the SNR of the user of interest in the presence of strong interferers, ambient noise and non zero signature cross-correlation. Estimates of the G auxiliary vector, the scalar parameter k and the input autocorrelation matrix that appears in the MOE design are based on 128 received signal samples.

FIG. 8 shows a comparison of the detector of the present invention with detectors of the matched filter, decorrelator, MOE types, as well as with the ideal minimum output energy detector on the basis of bit error rate as a function of the SNR of the user of interest in the presence of weak interferers, ambient noise and non zero signature cross-correlation. Estimates of the G auxiliary vector, the scalar parameter k and the input

autocorrelation matrix that appears in the MOE design are based on 128 received signal samples.

Returning now to FIG. 5, filter means 105, comprises essentially a transversal filter structure as is known in the art, and which includes an L-tap delay line 501, wherein L is the number of chips in the signature, or the processing gain; multiplier means 505, and summing means 510.

In general, FIR filters are well known and understood in the art. The design of such filters is known and is described in such texts as Adaptive Filter Theory, 2nd Ed., by Simon Haykin, Prentice Hall 1991, hereby incorporated by reference.

The L-tap delay line consists of L delay elements 501 having outputs coupled to a corresponding number of multiplier means 505. The L-tap delay line operates to sequentially shift, at the chip rate, sampled outputs($r_1 \ldots r_L$) during each signature interval. At the end of each signature interval, multiplier means 505 multiply the output of delay elements 501 with multiplier inputs, or tap coefficients ($w_1 \ldots w_L$). Tap coefficients ($w_1 \ldots w_L$) are provided by multiplier input generating means 407.

Filter means 105 may be implemented in hardware such as by flip flops, shift registers, or other known hardware devices as are conventionally employed in the art. Alternatively, filter means 105 may be implemented as a software filter wherein all of the input data samples are storable in memory. Several digital signal processors implementing digital filters are commercially available from a number of companies, including Texas Instruments. In addition, code listings for computer program implementations of the embodiments of the present invention are contained in Appendix B.

Software filter routines may perform the task of accessing a number of sample segments of the input data space, performing the calculations described herein, and

storing the resulting output sequence in an array of memory locations.

Summing means 510 sums the outputs of multiplier means 505 to provide summation means output 106. Summation means output 106 represents superposition of the multiplier outputs over one signature interval. Summation means output 106 is applied to sign determining means 107. The output of sign determining means 107 represents the symbol detected by the receiver.

Multiplier input generating means 407 provides multiplier inputs, or coefficients, designated $(w_1 \ldots w_L)$, as inputs to each of multipliers 505. Such coefficient generating means are not described in the prior art and are a significant feature of the present invention. Further details of the mathematical description of detector 500 including coefficient generating means 407 and filter means 105 is provided in Appendix A, and in a previously filed divisional application by the named inventors.

In a preferred embodiment, multiplier input generating means 407 includes digital signal processing means such as those available from a number of companies, including Texas Instruments, Inc.

Multiplier input generating means 407 is characterized by the set of multiplier inputs, or coefficients it generates. The set of coefficients is described by the vector $W = S_0 - kG$, where ${}_0S$ is the effective signature, or spreading code, of the user of interest. In a non multi path environment effective S_0 is the signature of the user of interest, while in a multipath environment effective S_0 is the multipath channel processed signature of the user of interest. G is a vector, referred to herein as an auxiliary vector, in R^L orthonormal to S_0, i.e.,

$\langle G,S_0\rangle = 0$, and $\langle G,G\rangle = 1$, where

$$G = \frac{E\{r_{\perp so}\}}{\|E\{r_{\perp so}\}\|}$$

and,

$$k = \frac{E\{<r, S_0><r, G>\}}{E\{<r, G>^2\}}$$

and wherein:

$$r_{\perp so} = sgn(<r, S_0>)\frac{r - <r, S_0>S_0}{\sqrt{\|r\|^2 - <r, S_0>^2}}$$

Alternate means for the generation of k is as follows:

$$k_{n+1} = k_n + \frac{c}{n} < r_n, S_0 - k_n G > < r_n, G >,$$

wherein n = 1,2,3,..., and c > 0 is an arbitrary scalar, or c/n = μ > 0 for all n, where μ is an arbitrary vector.

These relationships can be mathematically derived by the techniques and methods generally described in the previously filed provisional application relating to this invention and filed by the present inventors, and in Appendix A herein.

Multiplier input generating means 407, in a preferred digital signal processor (DSP) implementation stores the

bits representing S_0 in signature prestoring means 410 which may be a specified memory location in the memory area of the DSP itself. Input r comprising samples (r_1 ... r_L), shown at 420 are applied to multiplier input generating means 407 and filter means 105.

The signal representing an estimate of the transmitted symbol is applied to a sign determining means 107. Sign determining means 107 determines the sign, positive or negative, of the signal at its input and indicates the sign at its output. Any of a variety of well known components such as comparators may be employed in the present invention as sign determining means. The output of sign determining means 107 is a representation of the transmitted symbol.

An alternate embodiment of the demodulator of the present invention is implemented in a 2-tap Weiner reconstruction filter. Such a demodulator 600 is shown in FIG 9.

Input samples r_1, r_2, ...r_{L-1}, r_L, shown at 650, are applied one after another to subtracting means 615 where they are subtracted from the output of summing means 610. The difference signal is applied to adaptive control mechanism 625, described in more detail below.

Demodulator 600 has as input sequences $S_0[n]$ as shown at 601, and $G[n] + k\ S_0[n]$ as shown at 602. Input sequence 601 is the pre stored signature of the user of interest similar to that shown at 410 in FIG. 5. Input sequence 602 is generated by multiplier input generating means 623. Multiplier input generating means 623 also has as an input the sequentially applied input samples 650. Multiplier input generating means 623 generates sequence 602 based on the incoming samples and the signature and is determined according to the relationship $G[n] + kS_0[n]$, where G, k and S_0 are determined according to the formulas provided and described in the embodiment of FIG. 5, and n ranges from 1 to L. Input sequences 601 and 602 are applied to multipliers 605 and 607.

Adaptive control mechanism 625 generates multiplier inputs, or tap coefficients w_0, shown at 630; and $_1w$, shown at 635, by standard Least Mean Square (LMS) or Recursive Least Squares (RLS) algorithms known in the art and described in Adaptive Filter Theory, 2nd Ed., by Simon Haykin, Prentice Hall 1991, as incorporated previously herein by reference.

Multiplier inputs 630 and 635 are also applied to multipliers 605 and 607. The result is applied to second summing means 630 to produce a signal 660 representing an estimate of the incoming samples 650. The difference between the signal at 660 and the incoming sample at 650 represents the error in the estimate of the incoming sample that drives the adaptive control mechanism.

Sign determining means 612 determines the sign of 630 and thereby generates a representation of the transmitted symbol, or user information bit.

For vectors S_0 and G, and scalar k as previously defined, the mean square signal reconstruction optimal value of w_0 in Fig. 9 is:

$$w_0 = E_n\{\langle r, S_0 - kG\rangle\}.$$

The decision statistic w_0 is confined within a single symbol interval and the statistical expectation above is taken only with respect to the noise component n in the received signal r. The decision on the transmitted information bit is made by sign determining means 612 applied on w_0.

The filter of the present invention may be implemented according to methods well known and understood in the art. Using the teachings of the present invention, several options for implementing the present invention will be readily apparent to those skilled in the art.

These options include either off-the-shelf or customized hardware, the use of one or more programmable digital signal processors running programs which emulate the physical characteristics of a hardware version of the filter of the present invention, or a combination of hardware and software. Representative code listings and computer program algorithms for software implementations of the preferred embodiments of the present invention, including calculations for parameters S_0 ,k, and G are contained in Appendix B. All of these implementations remain within the scope of the present invention.

While the specification concludes with claims defining the features of the invention that are regarded as novel, the invention will be better understood from the following description in conjunction with the drawing figures, in which like reference numerals are carried forward.

APPENDIX A

I. Introduction and Background

Spread Spectrum communication systems have been successfully used in military applications for several decades. Currently, Direct-Sequence Code Division Multiple Access (DS/CDMA), a specific form of spread spectrum transmission, is receiving considerable interest in the race for establishing a new standard for digital mobile radio and personal communications systems [1]. Among other inherent advantages, CDMA endorsers expect better utilization of the available resources compared to the traditional multiple access schemes such as Time or Frequency Division. This may be particularly true in bursty user environments.

While the overall capacity of a CDMA system is determined by both the forward (base-to-mobile) and the reverse (mobile-to-base) link, most of the current research focuses on the reverse connection and deals with processing at the base station [2]. There, at the base station, we are allowed to make the -wishfully simplifying - assumption of known active user population (known *spreading codes* or *signatures*). However, having K active users simultaneously transmitting binary antipodal signals results in a 2^K-ary hypothesis testing problem to be solved in the R^L space, where L is the system processing gain or equivalently the dimensionality of the signature vectors. Even for moderate values of K and small processing gain L the computational requirements of the optimum multiuser detector are unrealistic [3]. On the other side of the spectrum, the handy conventional matched filter receiver (one per active user at the base station) suffers from unacceptable performance degradation in the presence of one or more high power multiple access (MA) interferers. This is the well documented "near/far problem" [3]-[4]. Consequently, matched filter receivers are only considered together with some form of stringent power

control that requires a separate feedback connection from the receiver to the transmitter [5].

These observations gave rise to a significant effort from the research community to develop suboptimal receivers with complexity linear to the number of users. It is expected that linear-filter receivers will provide a mid-way approach between the low-performance matched filter (MF) detector and the optimum multiuser detector. An important candidate within the class of suboptimal linear receivers is the so called *decorrelating detector*, originally proposed by Schneider [6] and analyzed and popularized by Lupas and Verdu [7]. In theory, the decorrelator achieves perfect multiple access interference (MAI) cancellation in the absence of any additive channel noise. In practice, MAI suppression comes at the expense of increased noise power at the output of the filter. In fact, the decorrelator can be viewed as a direct analog to the familiar zero-forcing equalizer for the single-user ISI channel [8]. Recently, a particularly elegant chip-rate adaptive implementation of the decorrelating receiver was proposed by Chen and Roy [9]. As a general comment, fast adaptive solutions are most promising for rapidly changing statistical environments such as mobile cellular systems.

A series of other suboptimal multiuser detectors have appeared in the pertinent literature and they all employ some form of successive MAI cancellation targeted on the stronger active users. Multistage detectors (Varanasi and Aazhang [10], [11] and Zie et al. [13]) fall in this category. A different approach with the same objective was followed by Duel-Hallen [8], [12], who considered decision feedback (DF) multiuser receivers. A completely different line of research involves the use of neural networks for multiuser receivers. A completely different line of research

involves the use of neural networks for multiuser detection. It was shown that both Perceptrons [14], [15], and Radial Basis Function (RBF) networks [16] can approximate arbitrarily closely the optimal decision boundaries. Not surprisingly, this comes at the expense of an exponentially large -with respect to the number of users- hidden layer of neurons. However, in all cases, reduced complexity neural networks trained in a supervised fashion gave satisfactory results. This approach (supervised learning) requires the use of a separate training sequence and a certain degree of coordination among the users. Arguably, this contradicts the very same nature of CDMA transmissions that allows for uncoordinated, independently acting users. In this context, adaptive *blind* receivers are more desirable. Reduced complexity blind (unsupervised) RBF designs were comprehensively studied by Mitra and Poor [16]. They report a certain level of performance degradation compared to their supervised counterparts, reduced near-far resistance, sensitivity on the initialization values, and "convergence times that may not be realistic for a real-time system".

The above discussion shows that multiuser detection has been the focus of the research community for several years. In cellular systems, multiuser detection is envisioned at the base station for the simultaneous recovery of all signals of the *known* intracell users [2]. On the other side of the coin we have the forward base-to-mobile link. In fact, the mobile user faces an even more challenging problem. Knowing only its own signature, the mobile is asked to detect its own binary signal in the presence of *unknown* spread spectrum interference and additive white Gaussian (AWG) channel noise. The theoretical difficulty of this problem, together with even tighter complexity, size, and weight requirements than the base station, has led to the

implicit or explicit assumption of the use of a MF receiver (with or without some form of power control [5]). It was only recently that some significant contributions of this crucial subject appeared in the literature. In [17] Madhow and Honig propose an L-tap adaptive FIR filter (L is the processing gain) that minimizes the mean-square error (MSE) between the output of the filter and the desired information bit. With proposed processing gain values L as high as 127[9], reduced length FIR filters were also examined. The need, however, for a separate training sequence remains. The MSE work was taken one step further by Honig et al. [18] and Madhow [19]. An L-tap FIR filter was blindly optimized in the minimum-output-energy (MOE) sense under the constraint that the energy of the user of interest is maintained. This receiver was shown to outperform the conventional MF detector by orders of magnitude. In theory, the blind FIR filter strikes the perfect balance between noise and Multiple Access interference suppression. Therefore, it is conceivable for this MOE receiver to outperform even the decorrelator which utilizes the assumed known interfering spreading codes. Its optimization, however, comes at a significant computational cost. The adaptive implementation (optimization) requires steepest descent in the R^L space. Again, with proposed values for L equal to 63 or 127 or higher, this is not a practical approach in general. The batch, closed-form, analytic solution asks for the inversion of the L x L sampled autocorrelation matrix of the received signal. In rapidly changing statistical environments such as a cellular system this costly inversion may have to be carried out very frequently. In addition it is expected that numerical instabilities encountered during the inversion of the *sampled* autocorrelation matrix may affect adversely the actual performance of the blind minimum output energy detector.

In this paper we deal with the same problem and we develop low optimization cost alternatives to the MF detector for the forward link of DS/CDMA systems. We consider the issues of single-user detection in *unknown* spread-spectrum MAI and AWGN from the Wiener Reconstruction-Filter point of view [20]-[21]. This way we are able to derive simple blind Wiener filter receivers with trivial 2-tap chip-rate adaptive implementations. It is satisfying to observe that, as it turns out, these receivers form a desirable subclass within the overall constraint optimization problem formulated in [18] and [19]. However, the important characteristic of the newly proposed detectors is the fact that they are parameterized with respect to only a single *scalar* parameter. Therefore, blind filter optimization can be carried out in a simple straightforward manner. Receivers that maximize the output SINR (Signal-to-Interference-plus-Noise Ratio) are readily derived. The closed form solutions for the single scalar tuning parameter are simple and require no matrix inversion. Adaptive on-line tuning procedures are also developed. They are fast, low complexity, strongly consistent Stochastic Approximation recursions [22] on R. The receiver that maximizes the output Signal-to-Interference-plus-Noise-Ratio is an alternative to the Minimum Output Energy (MOE) detector (originally proposed in [18] and used in [19] for acquisition and demodulation) with drastically lower optimization complexity. It is shown that, at only a minimal increase of the overall computational cost, this receiver outperforms significantly the conventional MF detector. In fact, in terms of bit error rate performance it compares favorably to the decorrelating detector with similar near-far resistance, although the latter utilizes the assumed known MAI spreading codes. While the target application for the proposed new class

of receivers was originally the forward base-to-mobile link, the above mentioned performance characteristics may place them as candidates for the reverse link, too. This could be particularly true for communications scenarios where strong interference from *neighboring* cells is present (intercell interference) or for the case where not all of the intracell spreading signatures are available to the base.

The rest of the paper is organized as follows. In Section II a brief review of the basic formulation of DS/CDMA transmission systems allows us to present our notation. Then, in Section III we introduce and analyze the concept of Reconstruction-Filter receivers. Section IV presents the core result of this work. A 2-tap reconstruction filter motivates the auxiliary-vector receiver which is then "blindly" optimized both in closed form and adaptively. Numerical results and simulations presented in Section V compares the newly proposed receivers with the conventional MF, the decorrelating detector, and the MOE receiver. A few final conclusions are drawn in Section VI.

II. System Modeling and Formulation

While the developments in this work are particularly well suited to the case of *asynchronous* CDMA systems, we found morae than one compelling reason to present these results in the theoretical context of *synchronous* CDMA systems. First of all, it is well known and understood that every asynchronous system can be modeled as an equivalent synchronous one with higher *effective* multi-user (MU) population [23]. The reader may refer to [17] for another comprehensive treatment that shows that in the worst case scenario every asynchronous interferer acts as two synchronous ones. In our treatment, the MU population is considered totally unknown. Therefore, the effective or actual

number of interferers plays no role of either mathematical or practical significance in the developments herein. In addition, the synchronous or quasi-synchronous approach is not that unrealistic, at least not for the forward link [2]. For example, interesting proposals call for quasi-synchronous satellite [24] and microcell CDMA systems [25]. Last but not least, confining ourselves within the context of synchronous CDMA allows for a certain level of notational simplicity that may help the clarity and focus of the presentation.

Let us then consider a CDMA system where K users transmit synchronously over an AWGN channel. The continuous-time received signal after carrier demodulation and low-pass filtering is modeled as follows:

$$r(t)=\sum_{i}\sum_{k=0}^{K-1}\sqrt{E_k}\,b_k(i)\,s_k(t-iT)+n(t) \qquad (1)$$

where, with respect to the *k-th* user, E_k is the *received* energy, $b_k(i) \in \{-1, 1\}$ is the *i-th* information bit, and $s_k(t)$ is the signature (spreading code). T is the symbol (bit) period and $n(t)$ is the filtered channel AWGN. The signature of every user is actually composed of L spreading chips and it is of the form

$$s_k(t)\sum_{j=1}^{L} c_k(j)\,P_{T_c}[t-(j-1)T_c] \qquad (2)$$

where L is the so called system *processing gain*, $c_k(j) \in \{-1, 1\}$, $j = 1,\ldots,L$ are the assigned signature bits for the *k-th* user, and $P_{T_c}(t)$ is the spreading pulse with duration $T_c = T/L$. Without loss of generality the signatures are assumed to be normalized such that

$$\int_0^T s_k^2(t)\,dt=1,\forall k=0,\ldots,K-1. \qquad (3)$$

Since we consider synchronous transmission we can immediately drop the index i from expression (1) and we can concentrate on a single information bit interval of length T. After conventional chip-matched filtering and sampling at the chip rate $1/T_c$ we organize the L collected samples in the form of an L-dimensional vector as follows:

$$r[n]\overset{\Delta}{=}\int_{(n-1)T_c}^{nT_c} r(t)\,P_{T_c}[t-(n-1)T_c]\,dt,n=1,\ldots,L. \qquad (4)$$

The R^L discrete time version of (1) for $r(t)$ can now be written as

$$r=\sum_{k=o}^{K-1}\sqrt{E_k}b_k S_k+n. \qquad (5)$$

Before we comment on (5) we establish some basic notation: Bold variables denote vectors (in R^L unless otherwise stated) or matrices, $E\{\cdot\}$ defines the statistical expectation operation, $\mathbf{x}^T$ is the transpose of $\mathbf{x}$, and $\mathbf{1}$ denotes identity matrices. Then, the random vector $\mathbf{n}$ is assumed to be WG with autocorrelation matrix $E\{\mathbf{n}^T\mathbf{n}\} = \sigma^2\mathbf{1}_{LxL}$. Without loss of generality, let us say that the user of interest is user O. Then the only known quantity in (5) is the signature $\mathbf{S}_0$ and the objective is to decide the transmitted information bit b_o. In this context, we may find it convenient to write the MAI term as

$$I=\sum_{k=1}^{K-1}\sqrt{E_k}b_k S_k, \qquad (6)$$

which allows us to write (5) as follows:

$$r=\sqrt{E_o}b_oS_o+I+n. \quad (7)$$

Equation (7) bottom-lines the nature of the problem we consider in this work. Operating in the R^L vector space, we have to detect a binary antipodal signal in the presence of *unknown* spread-spectrum multiuser interference and additive white Gaussian noise. For this purpose we may only use the available vector sample r and our knowledge of the desired spreading code S_o. We are now ready to introduce and study the concept of Wiener reconstruction-type filter receivers for the problem at hand.

III. Linear Reconstruction Filter Receivers

To set the stage for later developments we begin this section with the design of a linear Mean-Square optimal (Wiener) reconstruction filter for the detection of a binary signal in unknown DS/CDMA interference and AWGN [29], [31]. The section will conclude with the unifying observation that both the matched filter (MF) and the decorrelating receiver can be viewed and implemented as a special case of Wiener reconstruction filters.

Adding to the notation introduced in Section II, let $< \cdot,\cdot >$ and $\| \cdot \|$ denote the inner product and the norm respectively (both in R^L unless otherwise specified). Then, we recall that according to our formulation S_o is the spreading code of the desired user with $< S_o, S_o >= 1$ and antipodal coordinates $S_o[n] \in \{-1/L^{1/2}, 1/L^{1/2}\}$ for $n = 1,2,\ldots,L$. For the moment our objective is to consider an orthonormal basis for the space R^L that includes S_o, say $\{S_o, G_1,\ldots,G_{L-1}\}$ [26]. Interference vectors I orthogonal to S_o are of little interest, and so is the subspace spanned by $\{G_1,\ldots,G_{L-1}\}$.

Then, for some arbitrary scalar $\mu \neq 0$ and for each G_i we define.

$$G_i^x \stackrel{\Delta}{=} G_i + \mu S_o, \; i=1,\ldots,L-1. \qquad (8)$$

We note that $\langle G_i^x, S_o\rangle = \mu, \forall i=1,\ldots,L-1$, and we propose to attempt linear Mean-Square (MS) estimation of the received signal r using as "input" sequences the vectors

$$S_o, G_1^x, \ldots, G_{L-1}^X,$$

In essence, we perform MS *reconstruction* of the received signal from the

$$S_o, G_1^x, \ldots, G_{L-1}^X,$$

hypothetical components and this is the reason the term "reconstruction filter" was coined originally. Fig. 1 shows the corresponding L-tap adaptive Wiener filter. The tap-weight vector $W^T = [\omega_0, \omega_1, \ldots, \omega_{L-1}]$ can be adapted on-line via the RLS or the LMS algorithm. The suggested decision statistic for the detection of the b_o information bit of user S_0 is the tap-weight ω_0 itself. Therefore, the proposed detector is

$$b_o = sgn(w_o). \qquad (9)$$

The analytic solution for the optimal weighting coefficients (including the weight of interst w_o) is well known and it is given as a solution to the Wiener-Hopf equations (also known as Yule-Walker equations [20]). Let us organize our input sequences in the form of L x L matrix **D** with rows S_o, $G_1, \ldots, G_{L-1}$, that is

$$D^T = [S_o, \; G_1, \ldots, G_{L-1}] \qquad (10)$$

In our case, due to the deterministic nature of the input matrix D, the Wiener-Hopf equations in a matrix formulation can be written as follows:

where the expectation operations is taken with respect to the noise component only. The following theorem identifies the MS optimal weighting coefficient w_o. It also shows that the detector proposed in (9) possesses an interesting self MAI-cancellation property. The proof can be found in the Appendix. Theorem 1 *Let S_o and $G_1^*, G_2^* \ldots G_{L-1}^*$ defined by (8) for some u = 0 be vectors in the RL space used as input sequences for the Wiener reconstruction of the received signal vector r in (5) as shown in Fig. 1. Then,*

(i) the optimal weighting coefficient w_o is

$$w_o = E_n \langle r, S_o - u\sum_{i=1}^{L-1} G_i \rangle, \qquad (12)$$

and

(ii) for any instance of the received signal r, the filter S_o - u cancels completely all MAI vectors I within the interference subspace V1 spanned by

$$G_1^*, \ldots, G_L-1^*$$

A few important observations are now in order. First let us revisit (7). Part (ii) of Theorem 1 offers a strong deterministic-type result that guarantees for our proposed receiver perfect cancellation of all interference vectors I in V_I. This is very much in parallel to the Zero-Forcing property of the decorrelating receiver [6],[7] and the single-user ISI channel equalizer [8]. The space V_I is a function of the

parameter μ and given any interference vector I there exist μ such that I ϵ V_l (μ =) is the degenerate case <I. S_o> = 0). HOwever, as it is assumed that I is unknown, additional efforts are needed to determine an appropriate μ. In the next section we will reconsider this issue in detail.

Let us now investigate the meaning of the expectation operation in (12). Since our sole objective is the detection of the information bit b_o, we must confine ourselves and the adaptive Wiener filter within the corresponding symbol interval. Therefore, the expectation in (12) is taken only with respect to the AWGN vector n in (5) or (7), while E_o, b_o and I are all treated as unknown fixed deterministic parameters. In this context, Maximum Likelihood estimation of the expected value in (12) suggests simple sample-averaging [27]. Given the chip-matched sampling operation in (4) (at the chip rate $1/T_c$), we have only one sample vector r available. For w_o we then have to use a single-sample estimate:

$$w_o=\langle r,S_o-u\sum_{i=1}^{L-1}G_i\rangle, \tag{13}$$

Direct fractionally spaced oversampling of the baseband $r(t)$ in (1) is another possibility that is not considered in this present work. From expressions (7) and (13) we see that

$$w_o=\sqrt{E_o}b_o+\langle I,S_o-\mu\sum_{i=1}^{L-1}G_i\rangle+\langle n,S_o-\mu\sum_{i=1}^{L-1}G_i\rangle \tag{14}$$

and the noise power (variance) at the output of the receiver is

$$E\{\langle n, S_o-\mu\sum_{i=1}^{L-1} G_i\rangle^2\} = \sigma^2\|S_o-\mu\sum_{i=1}^{L-1} G_i\|^2. \qquad (15)$$

Before we conclude this section we will attempt to tie the newly proposed class of Wiener reconstruction filter receivers with existing CDMA literature. Let us consider the *single-tap* Wiener reconstruction filter with input sequence S_o. For this simple case (11) gives immediately $w_o = E_n\{<r, S_o>\}$ and dropping the expectation (single-sample estimate as explained before) we obtain the matched filter receiver. We just showed the following proposition:

Proposition 1 *The matched filter receiver* $< r, S_o >$ *is the result of optimal* single-tap *Wiener reconstruction of the received signal vector* r *with input sequence* S_o.□

Let us now change slightly the input sequences to the *L*-tap Wiener reconstruction filter of Fig. 1. In fact, instead of the sequences G_i of (8) let us use directly the orthonormal vectors G_i, $i = 1, \ldots, L-1$. In essence, we try to perform MS rexonstruction of r from an orthonormal basis of R^L that includes S_o. Equivalently we perform MS reconstruction of the interference component I of (7) in the subspace spanned by $G_1, G_2, \ldots. G_{L-1}$ which is orthogonal to S_o. It is well known that the receiver matched to S_o rejects all interference vectors orthogonal to S_o. Therefore, with this input set-up it is expected that the resulting MS optimal receiver is again the matched filter itself. Indeed, with $D^T = [S_o, G_1, \ldots, G_{L-1}]$ for this case, DD^T in

(11) becomes the identity matrix and we immediately obtain $w_o = E_n\{<r, S_o>\}$. This brief discussion can be viewed as motivation for definition (S). In a sentence, we should not use S_o in conjunction with sequences orthogonal to S_o if we wish to obtain something better than the matched filter receiver.

Finally, let us pretent for a moment that we do know the spreading codes of the other interfering users $S_1, S_2, \ldots, S_{K-1}$ in (5). In parallel to Fig. 1, let us consider the K-tap Wiener reconstruction filter with input sequences $S_o, S_1, S_2, \ldots, S_{K-1}$ correspondingly. As in (10) we organize our input in the form of a $K \times L$ data matrix, say D_o, with rows $S_o, S_1, S_2, \ldots, S_{K-1}$. Assuming linearly independent signatures, the solution to the Wiener-Hopf equations in (11) gives

$$W = (D_o D_o^T)^{-1} E_n\{[\langle r, s_o\rangle, \langle r, s_1\rangle, \ldots, \langle r, S_{K-1}\rangle]^T\},$$

(16)

where $D_o D_o^T$

is the $K \times K$ correlation matrix of the user spreading codes. Dropping the expectation (single-sample estimate) we see that the sign detector $b_i = sgn(w_i)$, $i = 0, \ldots, K - 1$, is the familiar multiuser decorrelating detector [6], [7]. In [9] Chen and Roy pursued an adaptive implementation of the decorrelating detector. They appraoched the problem from the Least Squares approximation point of view. Since Lease and Wiener reconstruction filters are statistically equivalent approaches, they obtained the same result (although their LS formulation required a more elaborate proof than the one presented above). Therefore, it is only

fair to credit the following proposition to Chen and Roy [9].

Proposition 2 *In a K-user CDMA environment, the decorrelating receiver is the result of optimal K-tap Wiener reconstruction of the received signal vector r with input sequences the user spreading codes* $S_{0,}$ $_S1,\ldots,$ K_{K-1}. □

Propositions 1 and 2 illustrate a different aspect of the familiar matched-filter and decorrelating receivers. However, the main contribution of this section is arguably Theorem 1. This theorem proposes a blind, chip-rate adaptive receiver with an inherent MAI cancellation property. Complexitywise, the size (length) L of the linear reconstruction filter may pose a problem. After all, this receiver is envisioned for the mobile user where we face tighter complexity, size, and weight requirements than the base station. Therefore, a careful consideration of lower size Wiener reconstruction filters appears well motivated. This is particularly true for multipath environments where RAKE-type processing requires one reconstruction filter per resolvable path. Section IV below submits our final proposal. It is a class of 2-tap linear reconstruction filters that can maximize the *average* output SINR. As we will see, the optimization of the 2-tap reconstruction receiver requires tuning of the cross-correlation parameter between the two input sequences.

IV. Auxiliary-Vector Maximum SINR Receivers

In this section, we use the theory of Wiener reconstruction filters as a vehicle for the derivation and presentation of a simple linear receiver that maximizes the output Signal-to-Interference-plus-Noise-Ratio (SINR).

Following the notation introduced earlier, let S_o be the spreading code of the user of interest and let G in R^L be an arbitrary vector orthonormal to S_o, that is

$$\langle G, S_o \rangle = 0 \wedge \langle G, G \rangle = 1. \tag{17}$$

Then, for some scalar $\mu = 0$, we consider the 2-tap Wiener reconstruction filter of Fig. 2 with input sequences S_o and G + __S_o. The proposed detector is again $b_o = sgn(w_o)$ as in (9). The Mean-Square (MS) optimal value of the weighting coefficient w_o and the inherent MAI cancellation property of the detector $sgn(w_o)$ can be viewed as a direct corollary to Theorem 1: Corollary 1 *For vectors* S_o *and* G *as in (17) and scalar* $\mu = 0$

(i) the MS optimal value of w_o *in Fig. 2 is*

$$w_o = E_n\{\langle r, S_o - \mu G \rangle\},$$

and

$$\tag{18}$$

(ii) the linear filter $S_o - \mu G$ *cancels all MAI vectors I in the direction of* $G + \mu S_o$. □

As explained in detail in Section III, our decision statistic w_o is confined within a single symbol interval and the statistical expectation in (18) is taken only with respect to the noise component n in (7). Therefore, we drop the expectation, we use a single-sample estimate, and we write

$$b_o = sgn(\langle r, S_o - \mu G \rangle). \tag{19}$$

Certainly, the introductory result in part (ii) of Corollary 1 is far from satisfactory. However, we will see tht the linear filter $S_o - \mu G$ can be easily optimized in a blind fashion with respect tot the scalar μ for any preselected auxiliary vector G. While in Theorem 1 of the previous section we had no interest in the specifics of the orthonormal basis $\{S_o, G_1, \ldots G_{L-1}\}$, here it is of interest to have the *single* auxiliary vector G selected appropriately. To gain some insight and increase our intuition with respect to the selection of G, it is useful to write the output expression for our proposed filter. With input r given by (7) we have

$$\langle r, S_o - \mu G \rangle = \sqrt{E_o} b_o + \langle I, S_o \rangle - \mu \langle I, G \rangle + \langle n, S_o - \mu G \rangle \tag{20}$$

Purely theoretically, (20) shows that the zero-forcing (ZF) solution can still be achieved, for any G constrained by (17) with <I, G> = 0, if we set $\mu = \frac{\langle I, S_o \rangle}{\langle I, G \rangle}$. The geometric interpretation is shown in Fig. 3 where μ, acting as a sterring parameter, places the filter $S_o - \mu G$ orthogonally to the MAI vector I. However, the price paid for this theoretical ZF solution is an increase of the noise power at the output by $\mu^2 \sigma$. This can be seen from the expression for the *average* variance of the filter output

$$E\{\langle r, S_o - \mu G \rangle^2\} = E_o + E\{\langle I, S_o \rangle - \mu \langle I, G \rangle\}^2\} + (1 + \mu^2)\sigma^2, \tag{21}$$

where the expectation is taken across symbol intervals, with respect to the information bits b_i, $i = 0, 1, \ldots K - 1$, and n. To minimize this impact it appears that we must maximize <I, G> or in other words we have to select a vector G as close to I as possible. By construction

(cf. (17)). $<I,G> = E_n\{<r,G>\}$ and we may insead select a G as close to r as possible, taking always into account the constraints in (17). To pursue in the sequel the statistical optimization of the filter with respect to the steering parameter μ, we need to have G fixed across symbol intervals. Therefore, we take the above arguement in the mean sense and we let the auxiliary vector G be -within a sign ambiguity- the average (normalized) projection of r onto the subspace orthogonal to S_o. The sign of this projection can be either $sgn(<r.S_o>)$ or $-sgn(<r,S_o>)$. Therefore, without loss of generality, if we write

$$r \perp S_o = sgn(<r, S_o>) \frac{r - <r, S_o> S_o}{\sqrt{\|r\|^2 - <r, S_o>^2}} \qquad (22)$$

then

$$G = \frac{E\{r \perp S_o\}}{\|E\{r \perp S_o\}\|}$$

We also notice that the vector $r \perp S_o$ given by (22) leads to perfect the reconstruction in the context of the optimal Wioener filter of Fig. 2 (i.e. $\hat{r} = r$). We still need to select μ.

The expression for the *average* output variance in (21) shows that the receiver $S_o - \mu G$ that maximizes the *average* output Signal to Interference plus Noise ratio (SINR) is in fact the well known Minimum-Variance-Distortionless-Response (MVDR) filter [20]. Indeed, the $S_o - \mu G$ filter is "distortionless" in the S_o direction ($<S_o, S_o - \mu G> = 1$) for any $\mu = 0$ and G constrained by (17). Then, simple minimization of the variance expression in (21) with respect to μG supresses all other input components that are not in the S_o direction and leads to the maximum SINR receiver. This result is in perfect agreement with previous work on this subject

[18]-[19]. However, in contrast to any previous approaches we no proceed considering *scalar parameterized* linear MVDR receivers and we fix the auxiliary vector G as explained before. Let us now find the value of μ, say , that minimizes the variance $E\{<r, S_2 - \mu G>^2\}$. Under regularity conditions and setting the derivative of the variance with respect to μ equal to *0*, we can prove in a straightforward manner the following proposition.

Proposition 3 *If* S_o *is the spreading code of the user of interest and G is some auxiliary vector constrained by (17) then the value of the sterring scalar* μ *that minimizes the variance expression* $E\{<r, S_2 - \mu G>^2\}$ *(maximizes the average output SINR) is*

$$\mu MVDR = \frac{E\{<r, S_o><r, G>\}}{E\{<r, G>^2\}} \quad \square \qquad (24)$$

Now, with $\mu MVDR$ given by (24), simple algebraic manipulations lead to the following variance expression:

$$E\{<r, S_o - \mu MVDR\, G>^2\} = E\{<r, S_o>^2\} - \frac{E^2\{<r, S_o><r, G>\}}{E\{<r, G>^2\}} \qquad (25)$$

Summarizing the results presented in this section, the proposed linear maximum SINR receiver is given by (19) with G defined by (22) and (23) and μ given by (24). The steering parameter μ in (24) requires long-term averaging for any fixed G. Should se find more appropriate an on-line optimizaiton method, a Stochastic Approximation recursion for μ is readily available

$$\mu n+1 = \mu n + \frac{c}{n} \langle r_n, S_o - \mu_{nG} \rangle \langle r_n G \rangle, \qquad (26)$$

where $n = 1, 2, \ldots,$ and $c > 0$. The adaptive procedure is a fast -with one update per symbol interval-recursion of *R* that converges with probability 1 to the value in (24)[22]. If we wish to trade strong

consistency for quick reaction to statistical changes and convergence in the mean, we can use a carefully selected constant learning rate p > 0 instead of the monotonically decreasing *rate c/n*. Extensive studies on the learning rate subject can be found in [21] and [28].

In the next section, we present numerical and simulation comparisons of the newly proposed detector with the conventional matched filter, the Minimum-Output-Energy, and the decorrelating receiver. We recall that the latter assumes known spreading codes for all interfering users.

V. Numerical Results and Simulations.

We examine a scenario of four users each equipped with a signature of length $L = 15$ and signature cross-correlation matrix D_oD as in (16) given by

$$D_o D_o^T = \frac{1}{15} \begin{bmatrix} 15 & 11 & 11 & 11 \\ 11 & 15 & 7 & 7 \\ 11 & 7 & 15 & 7 \\ 11 & 7 & 7 & 15 \end{bmatrix} \tag{27}$$

We compare the Bit Error Rate (BER) performance of the proposed auxiliary-vector receiver with the conventional matched filter, the Minimum-Output-Energy, and the decorrelating receiver for synchronous CDMA transmission over an AWGN channel. The BER is analytically evaluated for all schemes. However, required ensemble averages (such as the autocorrelation matrix of the received signal for the MOE receiver, and the auxiliary vector G and the steering parameter k_{MVDR} for the proposed receiver) are replaced by sample average estimates based on a data record of 128 samples. The BER of the *ideal* MOE (that is with *perfectly* known inverse

autocorrelation matrix of the received signal) is also included as a reference point.

Figure 4 shows the BER of the user of intrest as a function of its SNR over the 0 to 15 dB range ($SNR_i = E_i/\sigma^2$ for user i = 0,1,2,3). The SNRs of the interfering users are fixed at 3dB, 5dB, and 6dB respectively.

Figure 5 shows the BER as a function of the SNR of the user of interest in the presence of strong MAI (with corresponding SNRs fixed a 9dB, 9.5dB, and 10dB).

In Figure 6 the SNR of the user of interest (user 0) is fixed while the SNRs of the interfering users are varied. The BER is plotted as a function of the near-far coefficient *NFC* (identical for all interferers) which is defined as the ratio of the SNR of each interferer over its initial value. The SNRs for the interfering users are set initially equal to 8, 9 and 10 respectively, and are varied linearly (not in dB) according to: $SNR_1 = 8 \times NFC$, $SNR_2 = 9 \times NFC$, and $SNR_3 = 10 \times NFC$.

Finally, in Figure 7 we compare the convergence rates of the MOE and the auxiliary-vector receiver. The SNR of the user of interset if 12dB and the SNRs of the interferes are 9dB, 9.5dB, and 10dB. The BERs are plotted versus the number of samples used for the estimation of the inverse autocorrelation matrix of the received signal (MOE) or the estimation of μ and G (auxiliary-vector receiver). Both structures are initialized at the conventional signature-matched correlator and the results presented are averages over 10 independent Monte-Carlo simulations.

VI. Conclusions.

We reconsidered the concept of multiuser detection for DS/DSMA communication systems from the point of view of Wiener signal-reconstruction filters. We identified

both the decorrelating detector and the signature matched filter receiver as a direct special case of Wiener signal reconstruction. Generalizing this result we proposed an L-tap Wiener adaptive receiver with a powerful inherent MAI canceling property. However, the size of the filter (L taps where L is the system processing gain) may restrict the practicality of this approach. In view of these observations the natural, low complexity outcome of this line of work is a linear, scalar parameterized, auxiliary-vector receiver. The conceptual and computational simplicity of this receiver promises some immediate practical utility. The optimization can be carried out easily in a variety of different ways. In this work we chose to develop a blind (unsupervised) solution that maximizes the output Signal-to-Interference-plus-Noise Ratio (SINR). In future work we will consider optimization in the minimum probability of error sense (non-least-squares supervised learning [30]).

The newly proposed blind auxiliary-vector receiver compares favorably, both complexity-wise and performance-wise, to the decorrelating detector [6], [7], although the latter utilizes the assumed known signatures of the interferers. This is because the blind maximum SINR criterion, in contrast to the "decorrelating" criterion, strives to achieve the perfect balance between MAI and channel noise suppression. The optinal near-far resistance of the decorrelator appears closely matched by the auxiliary-vector receiver over a wide range of realistic near-far ratios.

The auxiliary-vector receivers are a subclass of the linear FIR filters. Therefore theoretically, that is for analytically known inverse autocorrelation matrix of the received signal, the blind auxiliary vector

receiver becomes a suboptimal low optimization complexity structure compared to the MOE FIR filter [18]. However, in practice, the exceptional optimization simplicity of the auxiliary vector approach leads to improved bit error rate performance, too. The reason is the costly, numerically unstable $L \times L$ sample autocorrection matrix inversion required by the MOE receiver that the blind auxiliary vector receiver avoids.

In view of these results, the linear, blind, auxiliary-vector filter becomes a candidate for the receiver of choice for the forward link of mobile cellular DS/CDMA communication systems. On the other hand, a bank of blindly optimized auxiliary-vector filters may be deployed as the reverse-link, base-station receiver.

ACKNOWLEDGMENT

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APPENDIX

Proof of Theorem 1

The expression for the MS optimal weighting coefficient w_o in part (i) is derived as a solution to the Wiener-Hopf set of equations. Let the data matrix D be as in (10), and let

$$R = DD^T \qquad (28)$$

be the correlation matrix of the input data sequences. By construction, cf. (8), the elements of the *symmetric* $L \times L$ matrix R are

$$R[1,1] = 1, \quad R[i,i] = 1 + \mu^2, \quad i=2,\ldots,L \qquad (29)$$

$$R[1,j] = R[j,1] = \mu, \quad j = 2,\ldots,L \qquad (30)$$

$$R[i,j] = R[j,i] = \mu^2 \quad i,j\neq 1, \text{ and } i\neq j. \qquad (31)$$

Let us now multiply the first row of R by μ and subtract it from all other rows. Due to (29)-(31) we see that

$$\det\{R\} = 1, \qquad (32)$$

where det{.} denotes the determinant. For a treatment of the properties of determinants the reader may be referred to [26]. From the Wiener-Hopf equations in (11) we have

$$RW=E_n\{[\langle r,S_o\rangle,\langle r,G_1^*\rangle,\ldots\ldots,\langle r,G_{L-1}^*\rangle]^T\}, \qquad (33)$$

where W is the Lx 1 tap-weight vector. Let us replace the first column of R with $E_n\{[\langle r,S_o\rangle,\langle r,G_1^*\rangle,\ldots,\langle r,G_{L-1}^*\rangle]^T\}$ and call the resulting matrix R_o. Then, $w_o = \det\{R_o\}/\det\{R\}$ *and due to (32).*

$$w_o = \det\{R_o\}. \qquad (34)$$

Again, basic row-operations on R_o give the desired result in (12):(a) We multiply the first row of R_o by μ and we subtract from all rows, and (b) we multiply rows 2.3....,L by μ and we subtract each one from the top first row.

To prove part (ii) we need to show that

$$\langle I,S_o-\mu\sum_{i=1}^{L-1}G_i\rangle=0$$

for all I in the subspace V_1 spanned by vectors $\{G_1^*,$

$G_2^*, \ldots . G_{L-1}^*\}$. So, let I be an arbitrary vector in V_1. This means that there are real numbers $c_1, \ldots, c_{L-1}$, w.l.o.g. not all 0, such that

$$I=\sum_{j=1}^{L-1} c_j G_j^* . \tag{35}$$

Then, for I as in (35)

$$\langle I, S_o-\mu\sum_{i=1}^{L-1} G_i\rangle=\sum_{j=1}^{L-1} c_j\langle G_j^*, S_o\rangle-\mu\langle\sum_{j=1}^{L-1} c_j G_j^*, \sum_{i=1}^{L-1} G_i\rangle . \tag{36}$$

Due to the orthonormality of the set $\{S_o, G_1, \ldots, G_{L-1}\}$ and the *i definition of G_i^*, $i=1, \ldots . L-1$ in (8), (36) implies $\langle I, S_o - \mu\Sigma^{L-1}_{i=1} G_i\rangle = \mu\Sigma_j c_j - \mu\Sigma_j c_j = 0$. This completes the proof.

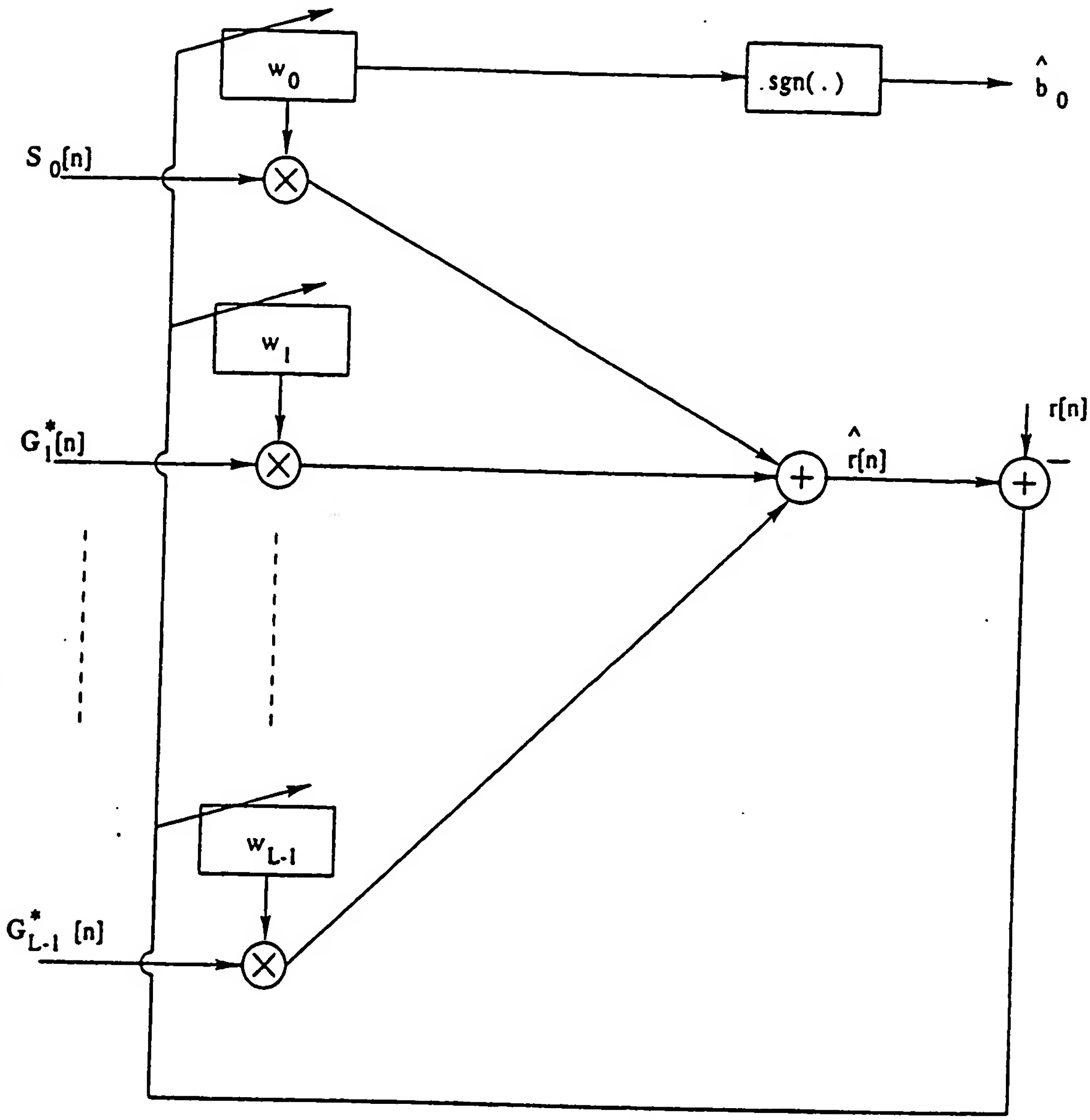

Figure 1: An Adaptive Wiener reconstruction filter receiver for the detection of the information bit b_o of user S_o.

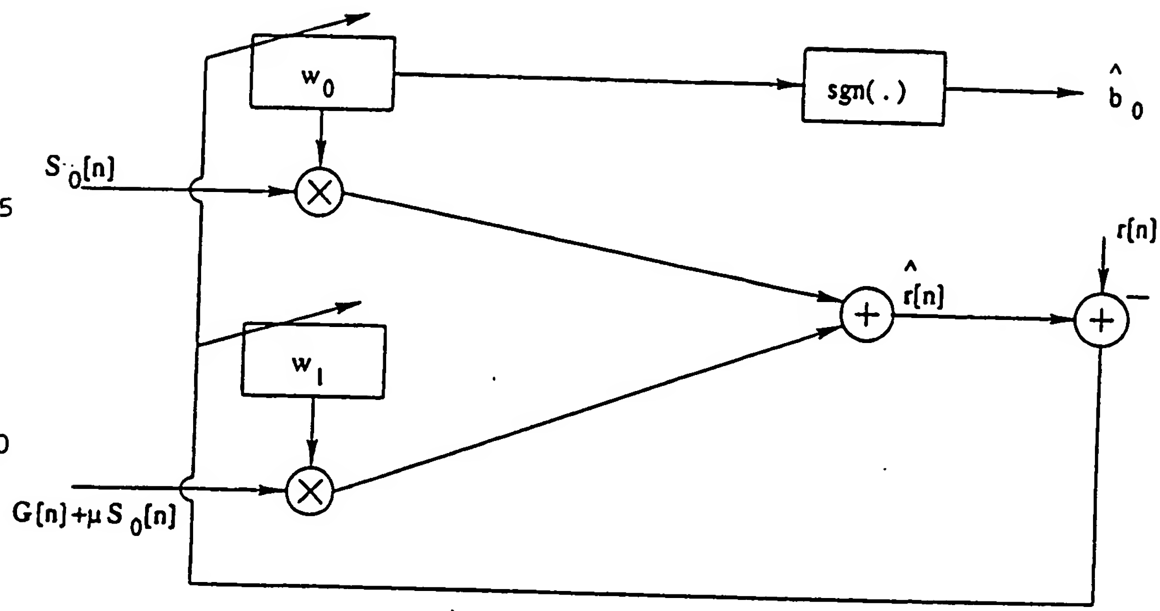

Figure 2: Two-tap Wiener-filter detector.

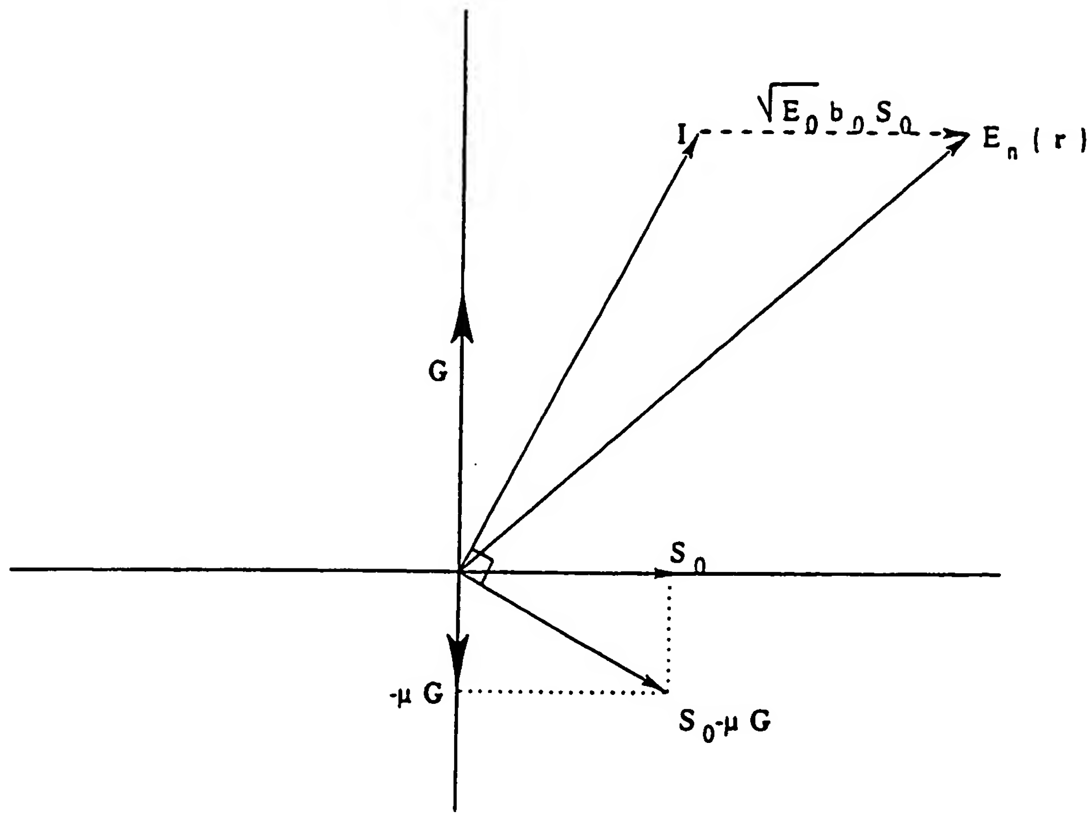

Figure 3: Geometric interpretation of the zero-forcing ability of the detector in Fig.2. The steering parameter μ places the filter $S_o - \mu G$ orthogonally to the MAI vector I.

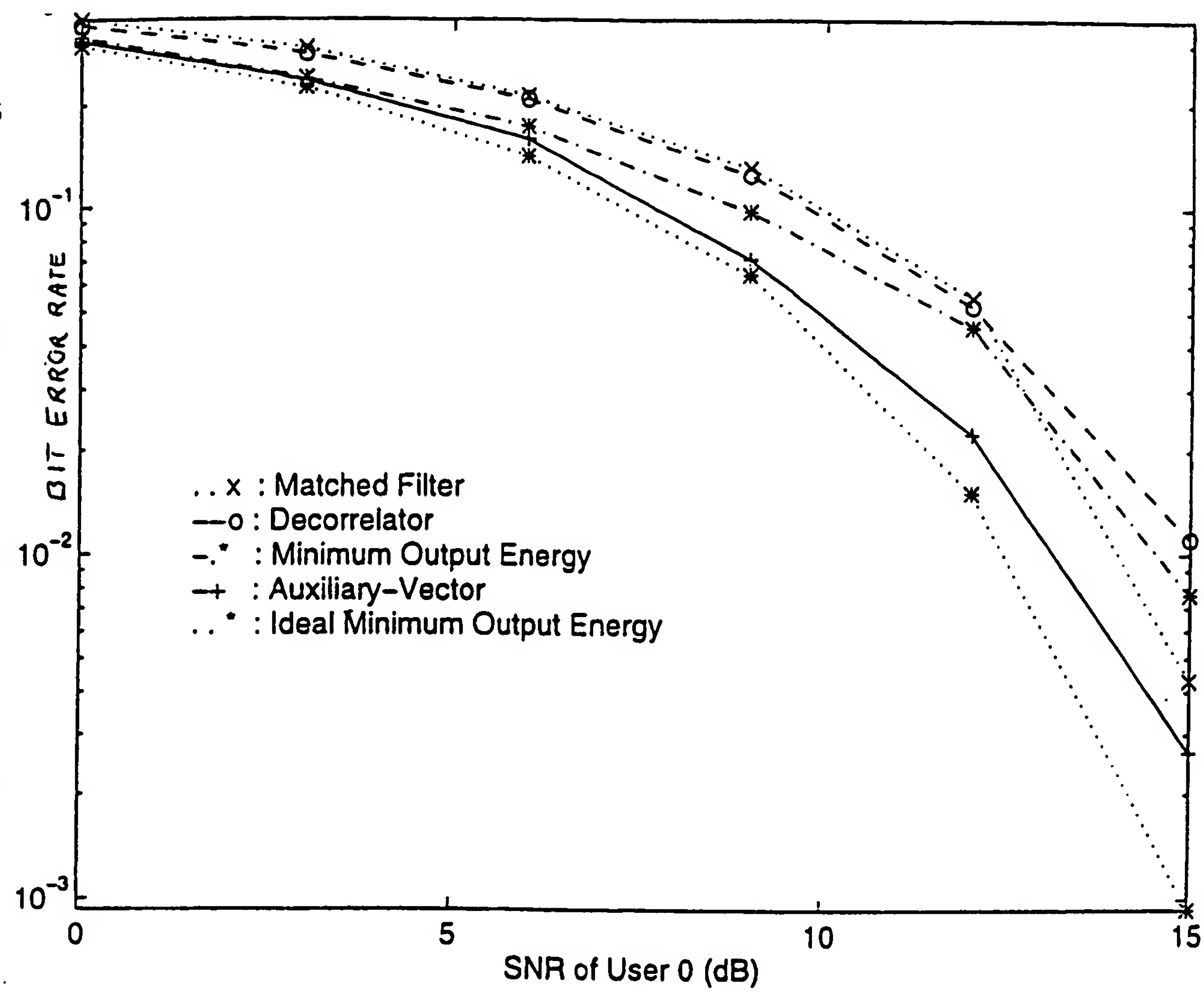

Figure 4: Bit error rate as a function of the SNR of the user of interest in the presense of weak interferers (SNR_1 = 3dB, SNR_2 = 5dB, SNR_3 = 6dB).

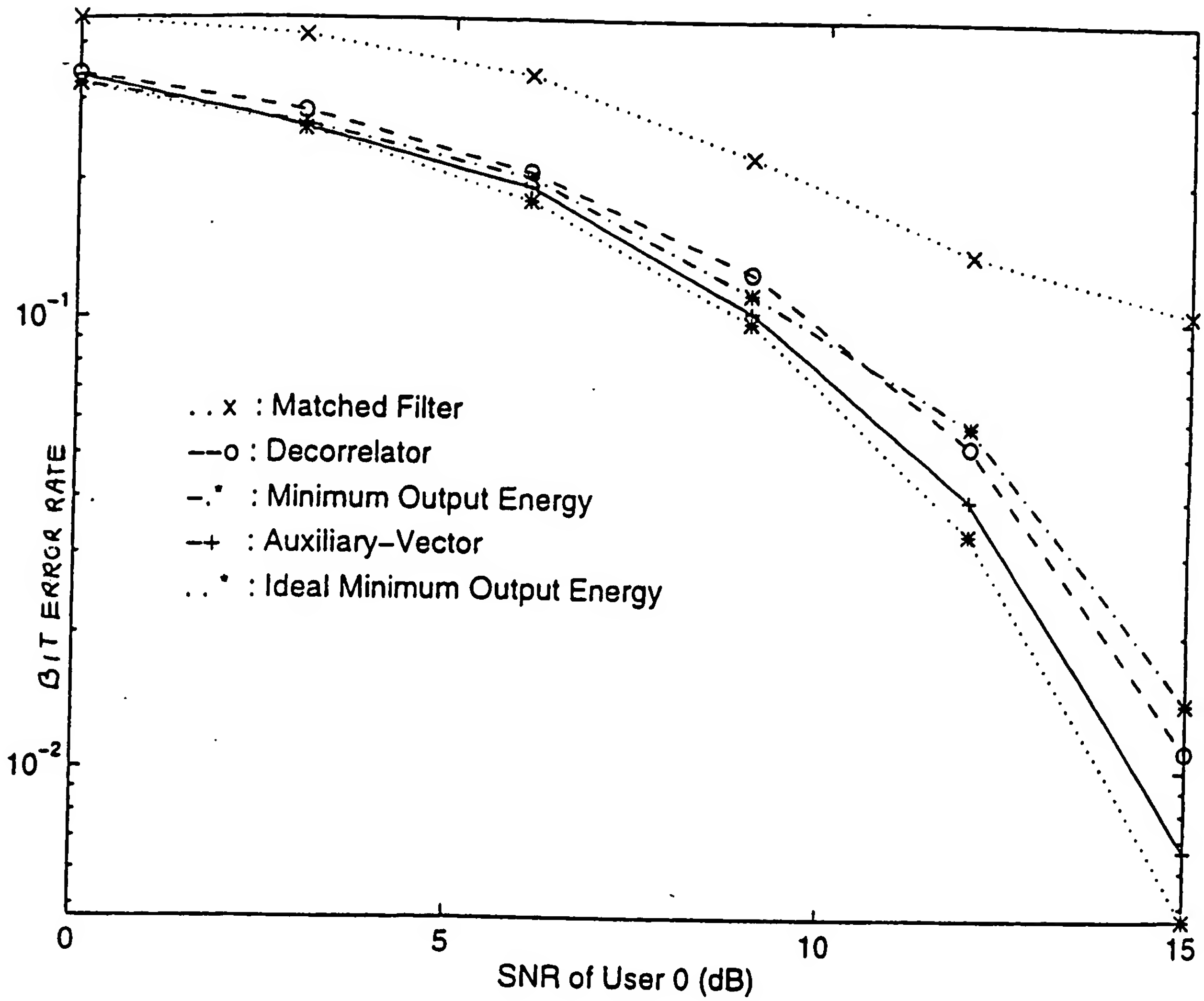

Figure 5: Bit error rate as a function of the SNR of the user of interest in the presense of strong interferers (SNR_1 = 9dB, SNR_2 = 9.5dB, SNR_3 = 10dB).

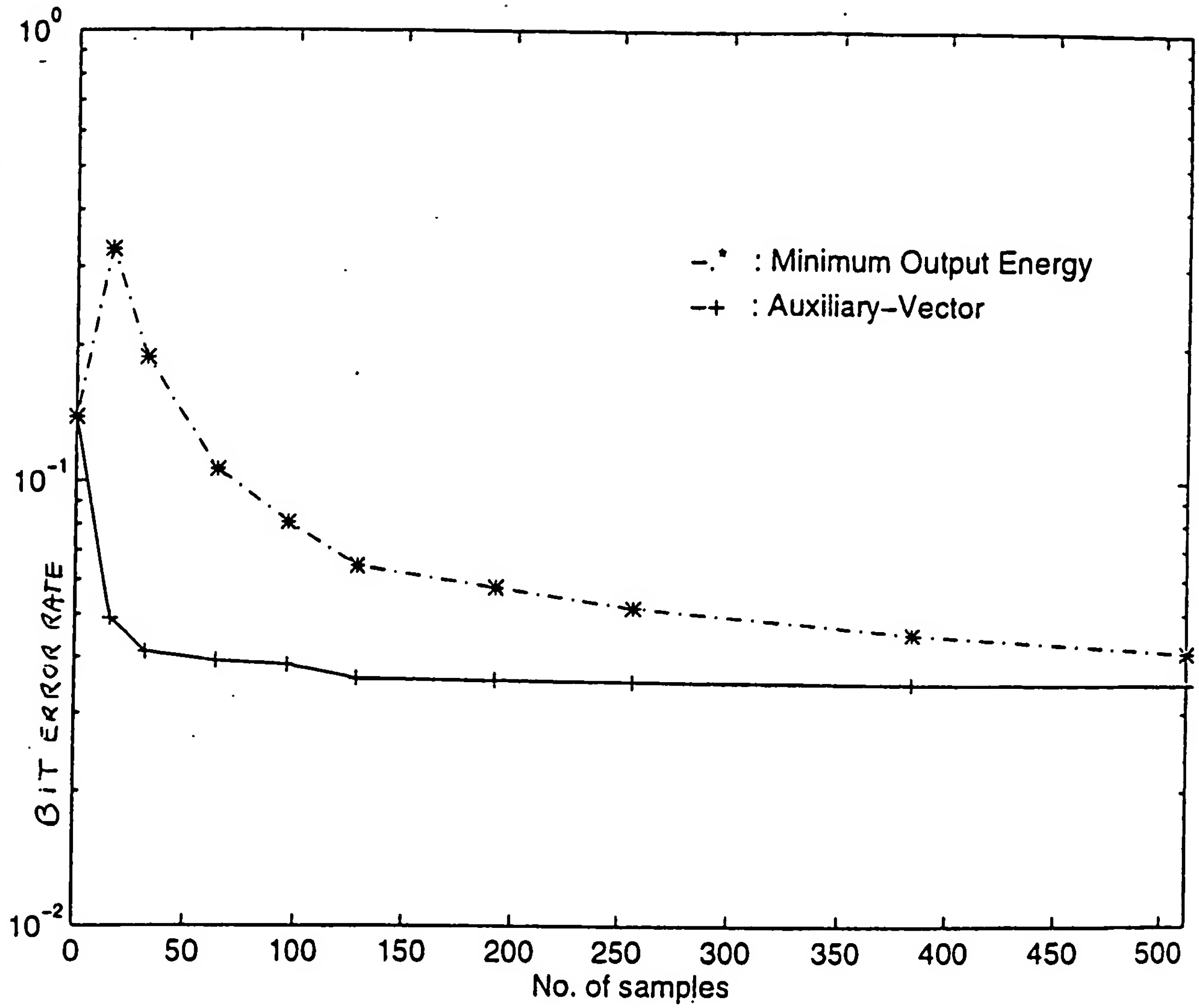

Figure 7: Bit error rate versus number of data samples for the MOE and the auxiliary-vector receiver (SNR_0 = 12dB, SNR_1 = 9dB, SNR_2 = 9.5dB, SNR_3 = 10dB).

APPENDIX B
CODE LISTINGS

```
function G=getG(r, E, S, sigma2,how)
% File: getG.m
% Input parameters:
% r: Matrix containing in its columns the received
vectors.
% E: The energies of the users.
% S: The signatures of the users.
% sigma2: The variance of the channel noise.
% how: How to calculate the vector G. See below.
% This function calculates the vector G from the
received vectors contained in the
%matrix r. This calculation can be done in two ways:
First, by substituting the mean
%value of the numerator of (27) by the sample average of
expression (26) (which is
%found using the received vectors contained in r).
Second, by using expression (26) and
%(27) after we drop the expectation the expectation
%in (27). In this case it is assumed that matrix r
contains only one received vector.
%The second way for calculating G is used in the
randomized receiver.

STEPS=length(r(1,:));            %How many received vectors
                                 we can use for the
                                 calculation
                             % of the vector G with the mean
                             average method.
GAIN=length(S(:,1));             %The system processing
                                 gain.
USERS-length(S(1,:));            %The number of users.
BITCOMBS=^USERS;             %The number of different
                             combinations of the
                             %transmitted bits.
b=getb(USERS);               %All the combinations of the
                             transmitted bits.

G=zeros(GAIN,1);                 %We initialize the vector
                                 G to all zeros.
SO=S(:,1);                       %The signature of the
                                 first user (user of
                                 interest).
if strop (how, 'meanav')==1      %When is equal to 'meanav'
                                 we find the vector
                             % G by calculating the mean
value of the expression (26)
                             % taking the mean average of
the expression using the
                             % received vectors in r.
     for I=1:STEPS
          dot=r(:,I)´*SO;        %The inner product of the
received vector and SO.
          G=G+sign(dot)*(r(:,I)-
dot*SO)./sqrt(r(:,I)´*r(:,I)-dot*dot);
```

```
        end
        G=G/STEPS;
    end
    if strop(how,'instant')==1     %When how is equal to
    'instant' we calculate G using
                                   % (26) and (27) after we drop
    the expectation in the
                                   % numerator and denominator of
    (27).
        dot=r´*SO;
        G=sign(dot)*(r-dot*SO)./sqrt(r´*r-dot*dot);
    end
    normG=sqrt(G´*G);              %we find the norm of vector G.
    G=G/normG;                     %And we normalize it.
```

```
function k=getk(r,E,S,G,sigma2,how);
% File getk.m
%Input parameters:
% r: Matrix containing in its columns the received
vectors.
% E: The energies of the users.
% S: The signatures of the users.
% G: The vector G of the auxiliary vector receivers.
% sigma2: The variance of the channel noise.
% how: How to calculate the steering parameter k. See
below.
% This function calculates the steering parameter k.
This calculation can be done in
%three ways: First, using expression (24) after we drop
the expectations in the numerator
%and the denominator. This way for calculating k is used
for the randomized receiver, and
%it assumes that the matrix r contains only one received
vector. The other two ways
%need the calculation of two expected values: one for
the numerator and one for the
%denominator of (24). In both ways, the following change
in expression (24) must be done:
%We must write the numerator as (SO´*r)*(r´*G), the
expectation of this is:
%SO´*E[r*r´]*G or SO´*R*G.
%Similarly the denominator must be written as G´*R*G.
%If we use for the calculation of k the form of
expression (24) there is a problem with
%numerical instabilities.
%Now, we have two choices for finding matrix R. We can
find it either analytically or
%by taking the mean average of the products of the
received vectors.  (See function getR).
%If k is calculated analytically the input matrix r is
not used.

SO=S(:,1);            %The signature of the first user
(user of interest).

if stromp(how, 'instant')==1  %If how is equal to
'instant' then to find k
                              % we use expression (24)
without the expectations.
     k=((r´*SO)*(r´*G))/((r´*G)*(r´*G));
end
if stromp (how, 'meanav')==1  %If how is equal to
'meanav' then to find k
                              % we use the modified
expression (24) where the
                              % matrix
                              % is calculated by taking the
mean average of the
```

```
                                % products of the received
vectors contained in r.
     R=getR(r,E,S,sigma2, 'meanav');
     k=(SO´*R*G)/(G´*G;
end
if stromp (how, 'analytic')==1     %if how is equal to
'analytic' then to find k
                                % we use the modified
expression (24) where the matrix
                                % is calculated analytically.
     R=getR(r,E,S,sigma2,'analytic');
     k=(SO´*R*G)/(G´*R*G);
end
```

```
function R=getR(R,E,S,sigma2,how)
% File getR.m
% Input parameters;
% r: Matrix containing in its columns the received
vectors.
% E: The energies of the users.
% S: The signatures of the users.
% sigma2: The variance of the channel noise.
% how: How to calculate the autocorrelation matrix R.
See below.
% This function calculates the autocorrelation of the
vectors contained in
% the columns of the matrix r.
%This calculation can be done in two ways: Either taking
the mean average of the
% product of the received vectors, or analytically by
taking the mean average
% of the products of the 2^USERS received vectors that
correspond to the
% different combinations of transmitted bits and adding
the autocorrelation matrix
% of the noise.


STEPS=length(r(1,:));      %How many received vectors we
can use for the calculation
                           % of the autocorrelation matrix with
the mean average method.
GAIN=length(S(:,1));       %The system processing gain.
USERS=length(S(1,:));      %The number of users.
BITCOMBS=2^USERS;     %The number of different
combinations of the transmitted bits.
b=getb(USERS);        %All the combinations of the
transmitted bits.

R=zeros(GAIN,GAIN); %We initialize the autocorrelation
matrix to all zeros.
if stromp(how, 'meanav')==1   %When how is equal to
'meanav' we must calculate the
                              % matrix R by taking the mean
average of the products
                              % of the received vectors which
are contained in r.
     for I=1:STEPS
          R=R+r(:,I)*r(:,I)';
     end;
     %Now, R contains the sum of the products of all the
received vectors.
     R=R./STEPS;     %We find its mean value.
end
if stromp(how,'analytic')==1  %When how is equal to
'analytic' we must calculate the
                              % matrix R analytically.
```

```
%For each one of the possible combinations of
transmitted bit we calculate the received
% vector with no additive noise. Using these vectors we
find the autocorrelation matrix
% and we add to it the autocorrelation matrix of the
additive white gaussian noise.
     for I=1:BITCOMBS
         bits=b(I,:);
         r=getr(bits,E,S,0);
         R=R+r*r';
     end
     R=R./BITCOMBS;
     R=R+sigma2*eye(GAIN);    %eye(n) is the n-by-n
identity matrix.
end
```

```
function r=getr(bits,E,S,sigma2)
% File: getr.m
%Input parameters
% bits: The user bits to be used for the calculation of the received vector.
% E: The energies of the users.
% S: The signatures of the users.
% sigma2: The variance of the channel noise.
% This function creates a received vector.

USERS=length(bits); %The number of users.
GAIN=length(S(:,1));     %The system processing gain.
r=zeros(GAIN,1);         %We initialize the received vector to all zeros.
for I=1:USERS
    r-r+sqrt(W(I))*bits(I)*S(:,I); %We add each user's contribution.
end
r=r+sqrt(sigma2)*randn(GAIN,1);  %Finally we add the channel noise.
```

```
function b=getb(nb)
% File: getb.m
%This function creates all the different 2^nb
combinations of nb bits, where the digit
%' 0 is substituted by the digit -1, since the users
transmit antipodal signals.
% First, it coverts all the integers from 0 to 2^nb-1 to
their binary (0,1) form
% and stores the results in the matrix B (which has
dimensions (nb,2^nb)).
% Then it changes all the zeros in the matrix b to -1.
To do that, the matrix is
% first converted to a vector using reshape. Then, the
0s are found using find and their
% positions are stored in the array z. All the elements
which their positions are
% stored in z are made equal to -1. Finally, the vector
b is converted back to a matrix
% unsign reshape.

b=zeros(2^nb,nb);          %We initialize matrix b to all
zeros.
%We store in the rows of b the binary representations of
all the integers
% form 0 to 2^nb-1.
for I=0:2^nb-1;
    b(I+1,:)=num2bin(I,nb);
end

reshape(b,1,nb*2^nb);     %reshape (x,m,n) is a matrix of
dimensions m by n m whose
                          % elements
                          % are taken columnwise from x.
r=find(b==0);             %find (x==1) is an array containing
the positions of the
                          % elements
                          % of the array x that are equal to
1.
b(z)=-ones(1,length(z));
reshape (b,2^nb,nb);
```

```
function bin=num2bin(num,Nb)
% File num2bin.m
% This functions converts an integer to an array of length Nb that contains
%the binary representation of the integer. The lease significant is the remainder of
%the division of the integer by two. The next bit is the remainder of the division by
%two of the result of the previous division and so on.

for I=1:Nb
    bin(Nb-I+1)=rem(num,2);
    num=floor(num/2);
end
```

```
function q=Q(x)
% File: Q.m
% This function calculates the integral from x to
infinity of the Gaussian pdf.

q=(1/2)*erfc(x/sqrt(2)); %erfc is the complementary
error function.
```

We claim:

1. A circuit for demodulating a received signal wherein the received signal includes a transmitted symbol, unwanted interference and noise, the circuit including:

means for obtaining one or more discrete samples of the received signal;

means, responsive to said means for obtaining one or more discrete samples, for filtering out the unwanted interference and noise, including multiplying means having at least two sets of inputs, at least one set being determined by the relationship between the parameters S_o ,k, and G.

2. A circuit according to claim 1 wherein said multiplying means comprises one or more multipliers, each of said multipliers having two inputs, wherein one of said inputs determined by the relationship between the parameters S_o ,k, and G.

3. A circuit according to claim 1 wherein the relationship between S_o ,k, and G is given by S_o - Kg.

4. A circuit according to claim 1 wherein said filtering out means includes a Finite Impulse Response Filter.

5. A circuit according to claim 1 wherein the relationship between S_o ,k, and G is given by G + Ks_o.

6. A circuit according to claim 1 wherein said filtering out means is a two tap Weiner filter.

7. The circuit according to claim 3 wherein said multiplying means includes a number of multipliers, the number of multipliers corresponding to the processing gain of the circuit.

8. A circuit according to claim 1 wherein said means for filtering out unwanted interference and noise includes an output, the circuit further including means responsive to said output for determining the sign of

said output thereby generating a generally accurate representation of the transmitted symbol.

9. The circuit of claim 1 wherein said means for filtering out the unwanted interference and noise includes a digital filter.

10. The circuit of claim 9 wherein said digital filter is a software program in a computer.

11. The circuit of claim 9 wherein said digital filter is a programmable hardware processor.

12. The circuit of claim 9 wherein said digital filter is a dedicated integrated circuit.

13. The circuit of claim 1 wherein said one or more samples are obtained at a rate corresponding to the chip rate of the transmitted symbol.

14. In a CDMA communication system, a circuit for demodulating one or more received signals by means of one or more receivers, each of said one or more receivers having a pre-assigned CDMA signature sequence corresponding to a mobile subscriber, and generating a detected symbol which represents a transmitted symbol, wherein the system has a pre-determined processing gain and a concomitant chip determined by the processing gain and the symbol rate of the system, the circuit comprising:

means for storing the pre-assigned CDMA signature sequence;

means for sampling the received channel signal to produce an incoming sampled signal;

multiplier input generating means having at least one input responsive to said means for sampling the received signal and having at least one input in communication with said means for storing the pre=assigned CDMA signature sequence, said multiplier input generating means having an output determined by the relationship between the parameters S_0 ,k, and G.;

means responsive to said means for sampling the incoming channel signal, for filtering unwanted interference and noise from said incoming sampled signal, said filter means including one or more multiplying means, said multiplying means in communication with said output of said multiplier input generating means, said filter means further including a filtered signal output;

sign determining means responsive to said filtered signal output, for generating a signal representing a detected symbol.

15. The circuit according to claim 14 wherein the receiver is synchronized with the source.

16. The circuit of claim 14 wherein the received channel signal is a downlink channel signal.

17. The circuit of claim 14 wherein the received channel signal is an uplink channel signal.

18. In a CDMA communication system wherein both uplink and downlink channel signals are propagated, a circuit for demodulating a received channel signal by means of a receiver having a given CDMA signature sequence, said receiver generating a detected symbol which represents a symbol transmitted by a source, wherein the CDMA signature sequence corresponds to a mobile subscriber, and wherein the source and receiver are synchronized, wherein the system has a pre-determined processing gain and a concomitant chip rate determined by the processing gain and the symbol rate of the system, the circuit comprising:

means for storing the given CDMA signature sequence;

means for sampling the received channel signal at the chip rate to produce an incoming sampled signal;

multiplier input generating means having at least one input responsive to said means for sampling the incoming channel signal and having at least one

input in communication with said means for storing a preassigned signature sequence, said multiplier input generating means having an output determined by the relationship between the parameters S_0 ,k, and G;

filter means responsive to said means for sampling the incoming channel signal, said filter means including multiplying means, said multiplying means in communication with said output of said multiplier input generating means, said filter means further including a filtered signal output;

sign determining means responsive to said filtered signal output, said sign determining means generating a signal representing a detected symbol.

19. A method for demodulating a received signal wherein the received signal includes a transmitted symbol and unwanted interference and noise, the method including the steps of:

a) obtaining one or more discrete samples of the received signal;

b) filtering out the unwanted inference and noise from said discrete samples wherein the filtering out step includes the step of multiplying at least two sets of inputs, at least one set being determined by the relationship between the parameters S_0 ,k, and G.

20. The method of claim 19 further including the steps of providing the output of step "b" to assign determining means, determining the sign of said filtered output and generating a representation of the transmitted symbol based on the sign determined in the previous step.

21. In a CDMA communication system wherein both uplink and downlink channel signals are propagated, a method for demodulating a received channel signal by means of a receiver having a pre-assigned CDMA signature sequence corresponding to a mobile subscriber, said receiver generating detected symbol which represents a

transmitted symbol, the source and receiver being synchronized, wherein the system has a pre-determined processing gain and a concomitant chip rate determined by the processing tine and the symbol rate of the system, the method composing the steps of:

a) storing the pre-assigned CDMA signature sequence;

b) sampling the incoming channel signal at the chip rate to produce an incoming sampled signal;

c) supplying said incoming sampled signal to filter means and multiplier input generating means, said multiplier input generating means having an output determined by relationship between the parameters S_o, k, and G;

d) providing said output to a sign determining means;

e) determining the sign of said output to hereby generate the detected symbol.

22. The method of claim 21 wherein said filter means is a transversal filter.

23. The method of claim 21 wherein said transversal filter is a Finite Impulse Response Filter.

24. The method of claim 21 wherein said filter means is a Wiener filter.

25. The method of claim 24 wherein said filter means is a Wiener reconstruction filter.

26. The circuit of claim 18 wherein the downlink channel signal is propagated form a base station to a mobile user and wherein said incoming channel signal is the signal received by the mobile user.

27. The circuit of claim 18 wherein the uplink signal is propagated from a mobile user or a base station and wherein said incoming channel signal is the signal received by the base station.

28. The method of claim 19 wherein the steps are performed by a program running in a computer.

29. The circuit of claim 1 wherein said filter means is implemented by means of a combination of hardware filter means and software filter means.

30. The circuit of claim 1 wherein the noise includes ambient noise.

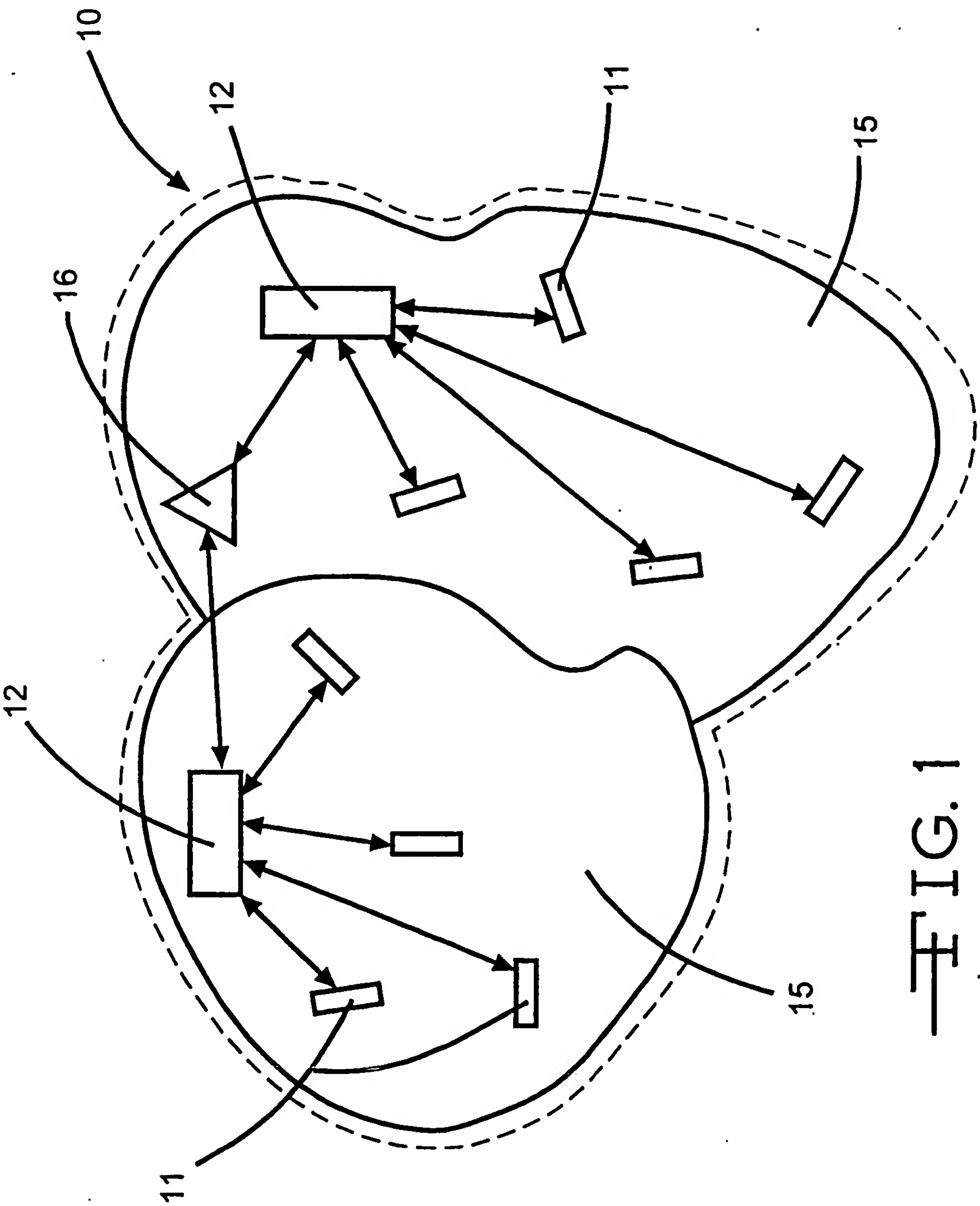

FIG. 1

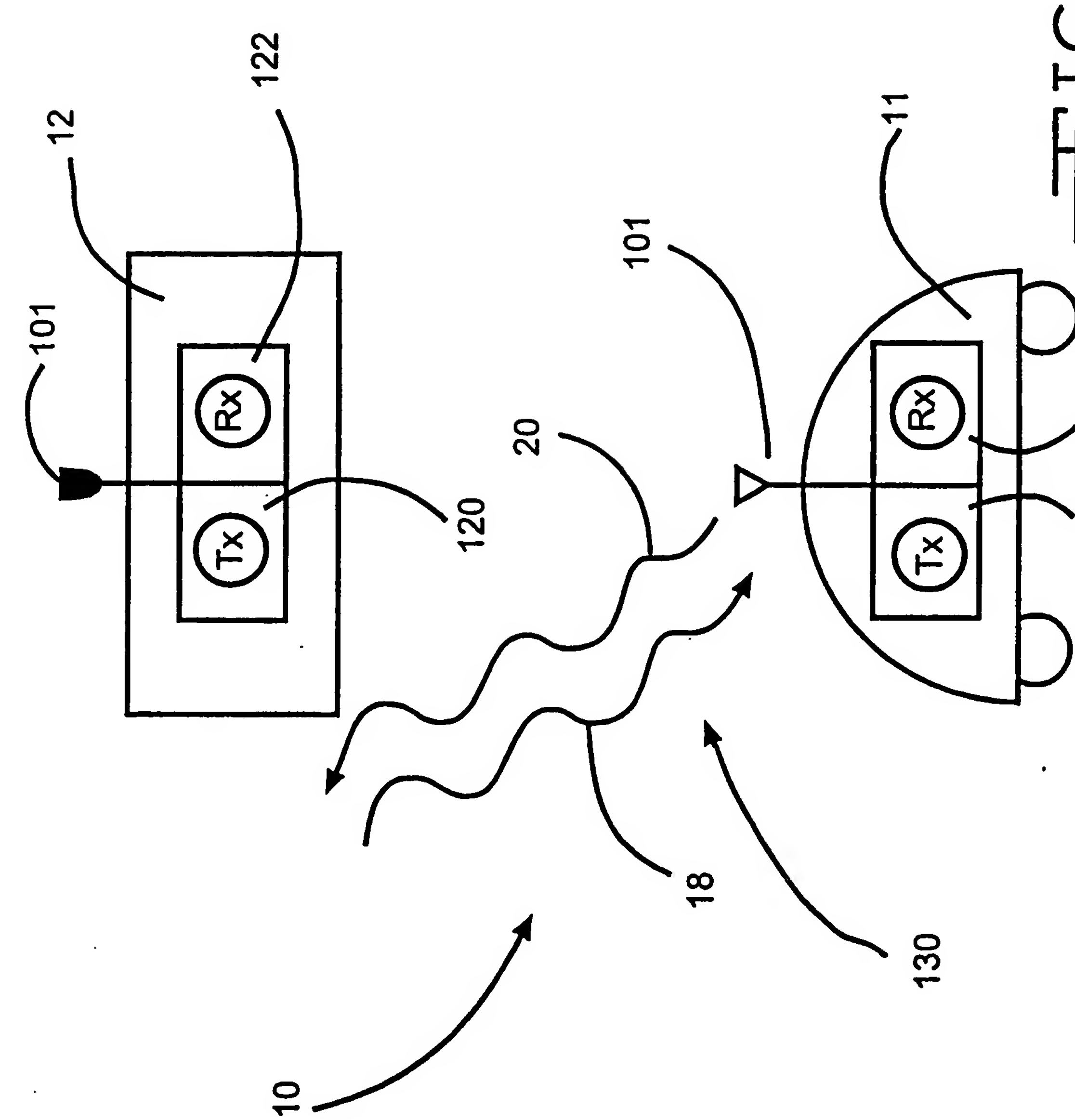

FIG. 2

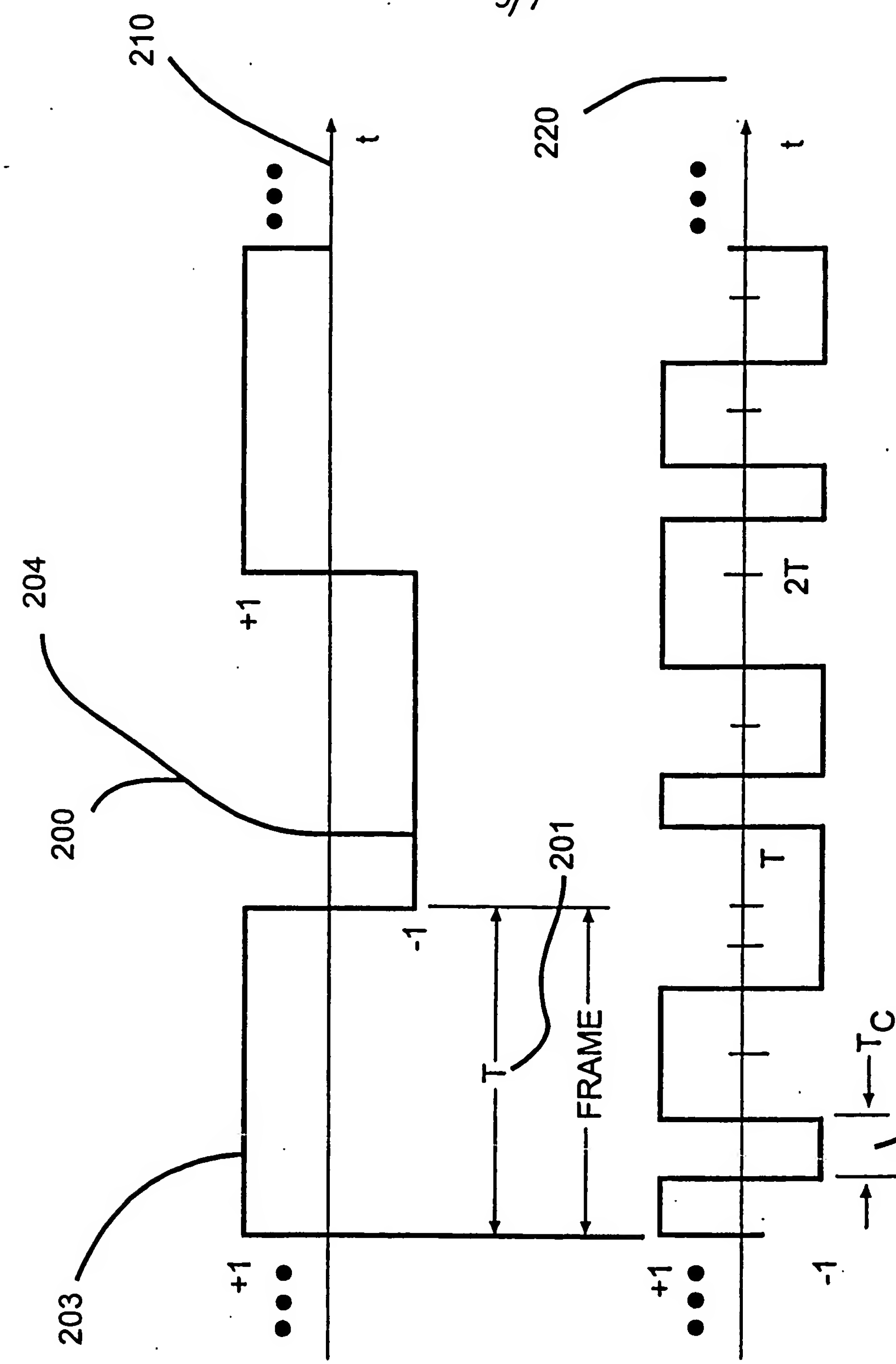

FIG. 3

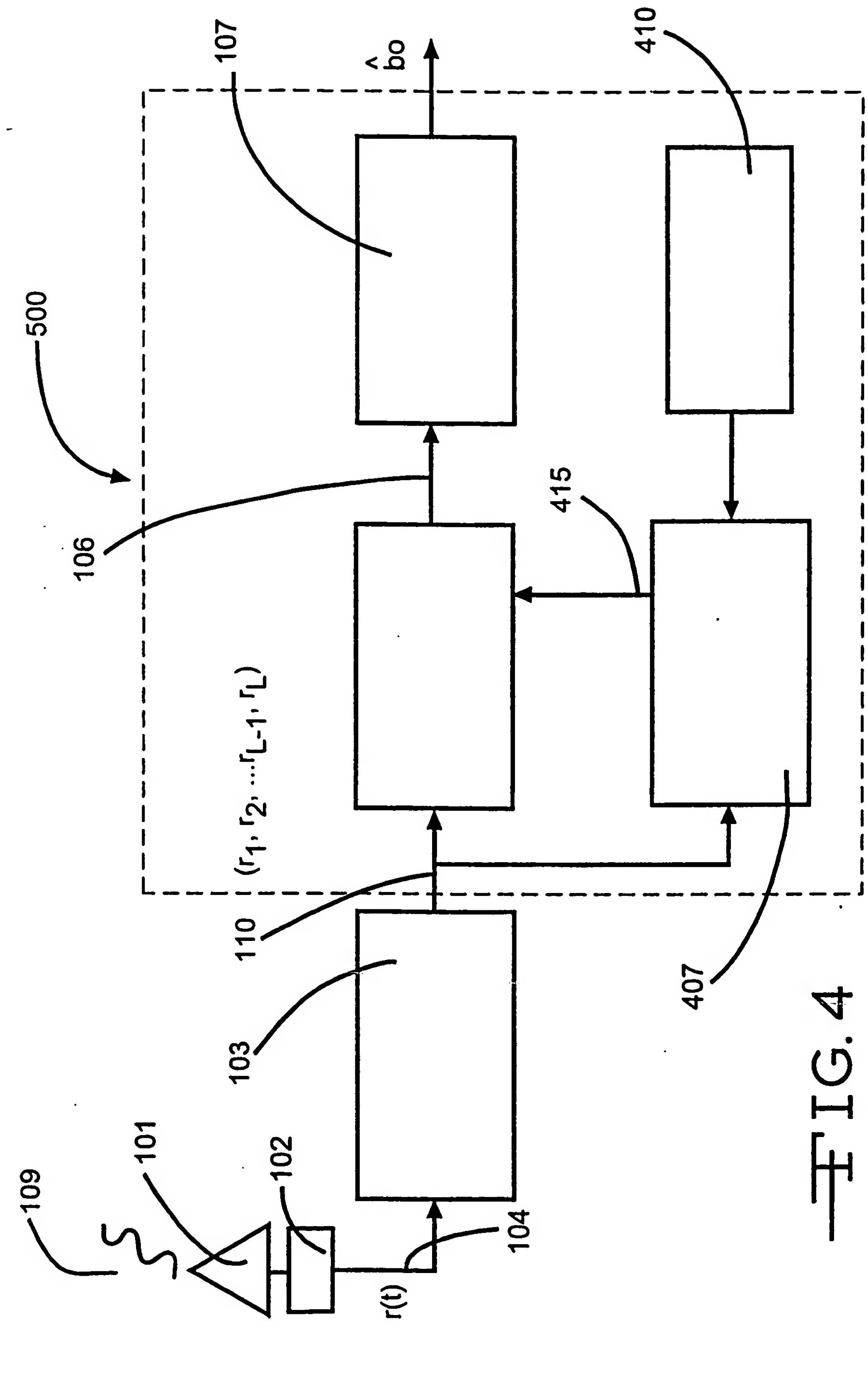
107
$\hat{b}_o$
410
500
106
415
$(r_1, r_2, \ldots r_{L-1}, r_L)$
110
103
407
109
101
102
104
r(t)

FIG. 4

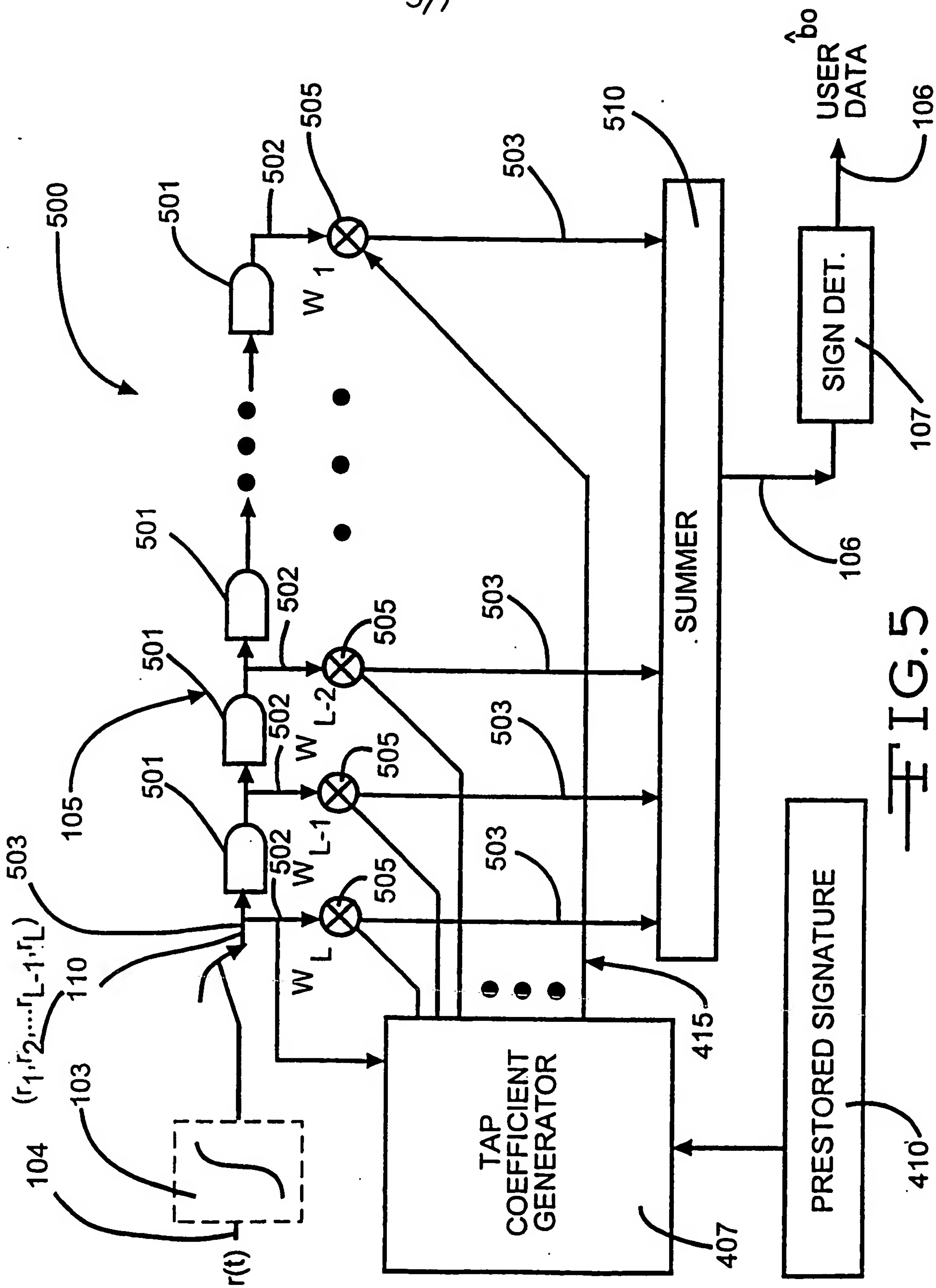
500
501
502
503
505
105
110
$(r_1, r_2, \ldots r_{L-1}, r_L)$
103
104
r(t)
W_1
W_{L-2}
W_{L-1}
W_L
510
SUMMER
SIGN DET.
107
106
USER DATA $\hat{b}_o$
TAP COEFFICIENT GENERATOR
407
415
PRESTORED SIGNATURE
410

FIG. 5

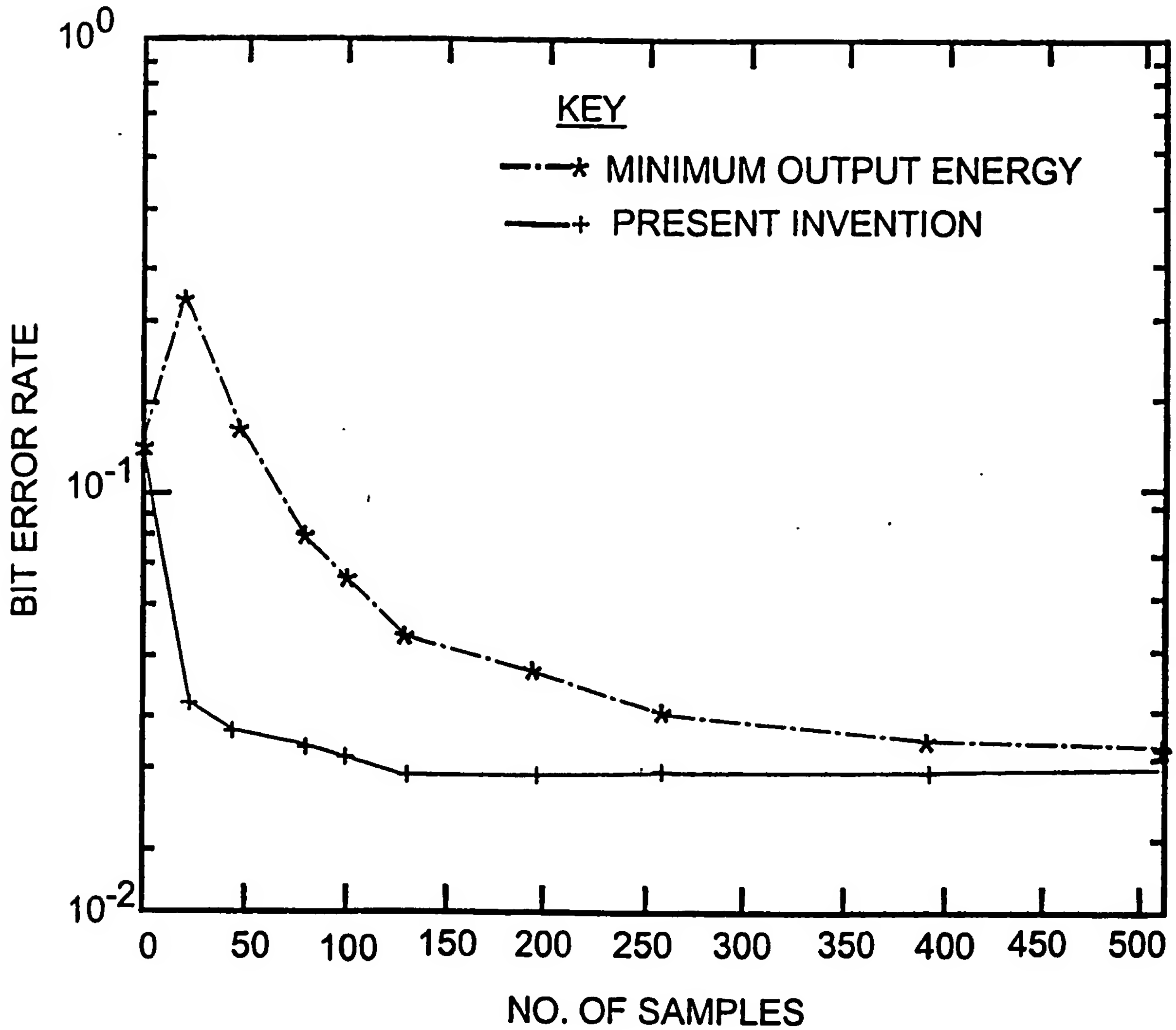
KEY
MINIMUM OUTPUT ENERGY
PRESENT INVENTION
BIT ERROR RATE
10^0
10^{-1}
10^{-2}
0
50
100
150
200
250
300
350
400
450
500
NO. OF SAMPLES

FIG.6

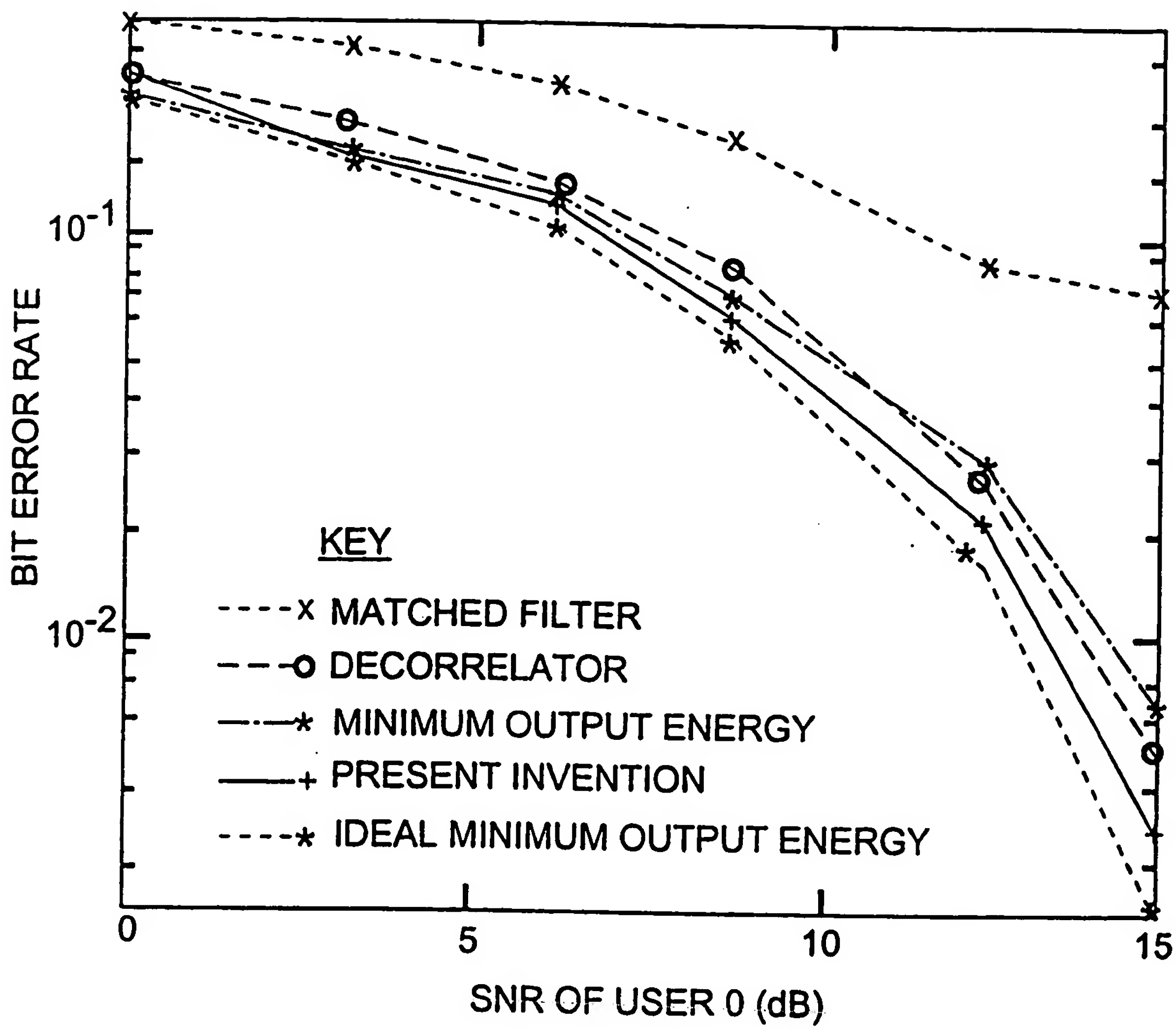
BIT ERROR RATE
10^{-1}
10^{-2}
0
5
10
15
SNR OF USER 0 (dB)
KEY
MATCHED FILTER
DECORRELATOR
MINIMUM OUTPUT ENERGY
PRESENT INVENTION
IDEAL MINIMUM OUTPUT ENERGY

FIG. 7

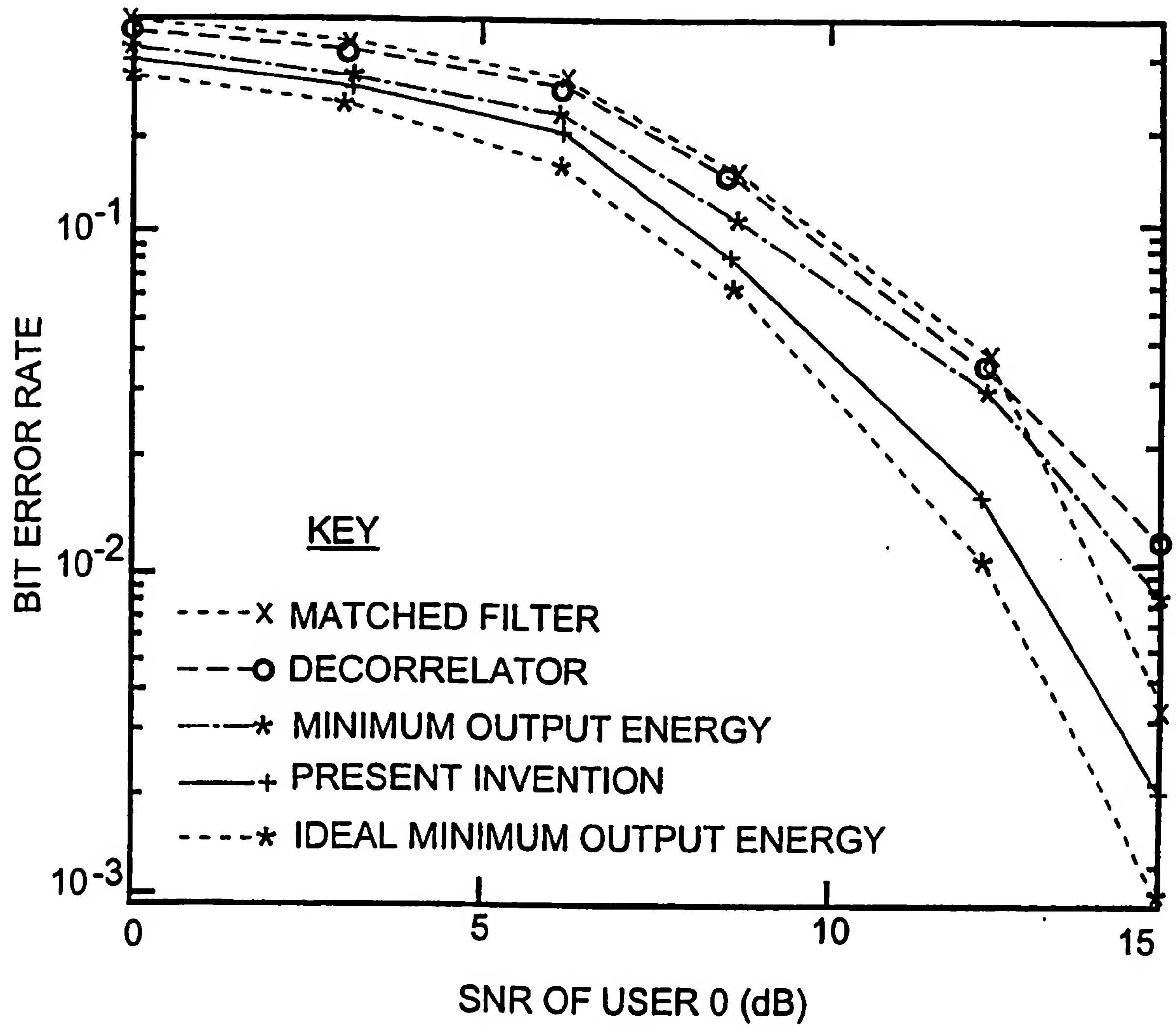
BIT ERROR RATE
10^{-1}
10^{-2}
10^{-3}
0
5
10
15
SNR OF USER 0 (dB)
KEY
MATCHED FILTER
DECORRELATOR
MINIMUM OUTPUT ENERGY
PRESENT INVENTION
IDEAL MINIMUM OUTPUT ENERGY

FIG.8

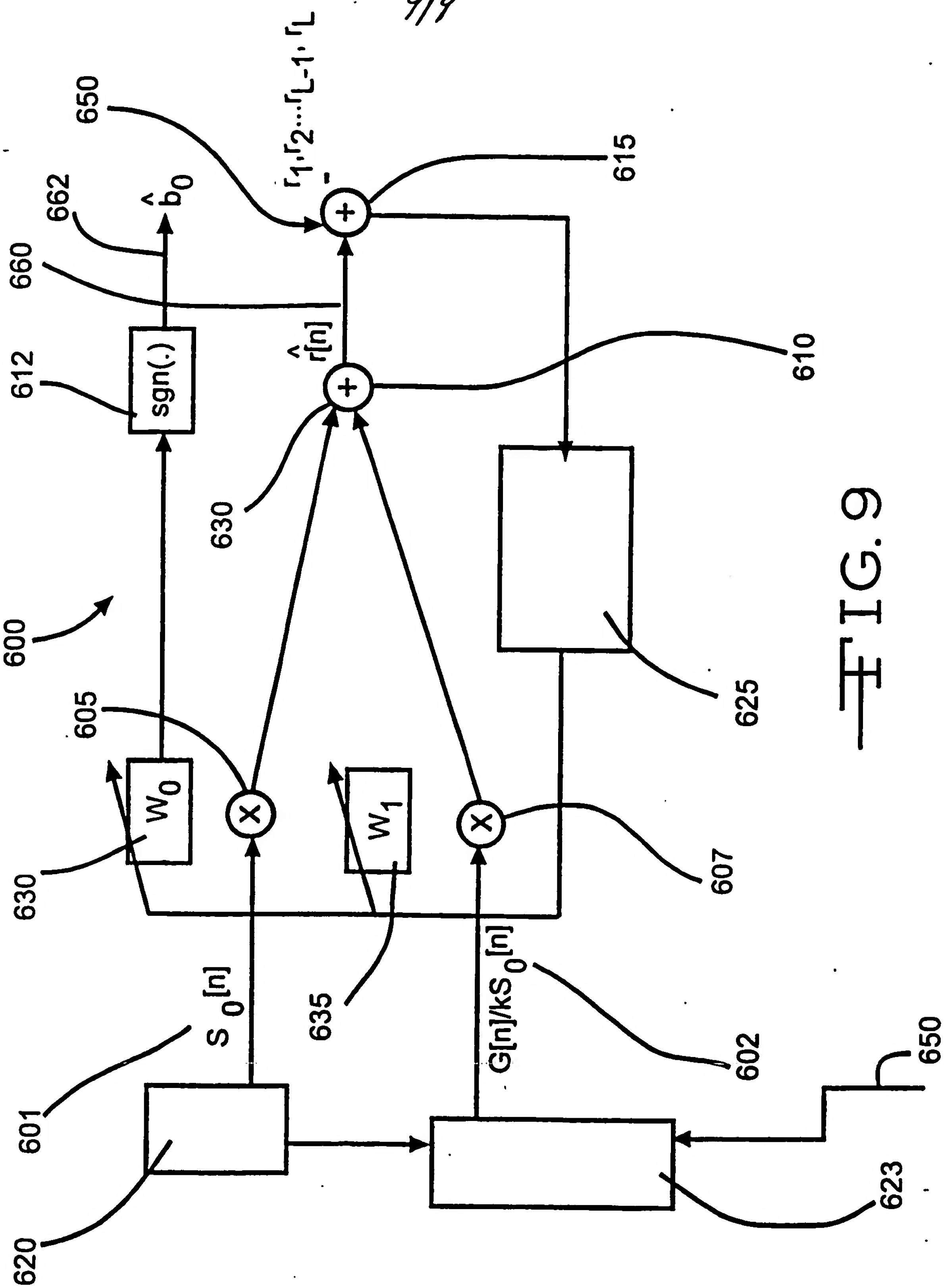
600
601
$S_0[n]$
602
$G[n]/kS_0[n]$
605
607
610
612
sgn(.)
615
620
623
625
630
630
635
650
650
660
662
$\hat{b}_0$
$\hat{r}[n]$
$r_1, r_2 \cdots r_{L-1}, r_L$
W_0
W_1

FIG. 9